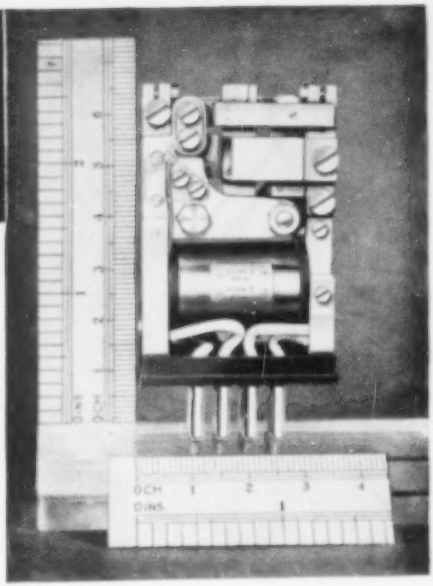
WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

APRIL 1950

VOL. XXVII TWO SHILLINGS AND SIXPENCE No. 319

CARPENTER POLARIZED RELAY TYPE 5



MAGNETIC polepieces constitute main frame and carry all other components. Hence no space is wasted on a frame structure, and great rigidity and high thermal stability is secured.

Axis of vibration of armature passes through centre of gravity, hence absence of positional error and virtual immunity from effect of vibration.

The only adjustment required is that of contact adjustment screws. No contact chatter.

Two generic types of the Relay are made, each of which may be had on either plug or solder tag base as follows :—

Dimensions with cover but excluding wiring tags:

HEIGHT	2 $\frac{5}{32}$ in.	56 mm.
WIDTH	1 $\frac{7}{16}$ in.	37 mm.
DEPTH	$\frac{3}{4}$ in.	19 mm.
WEIGHT	4.8 ozs.	137 gm.

TYPE 5A. Lightly side stable high sensitivity model, plug-in base, 12 pins. When adjusted to a static gap of .002 ins. this model has the following characteristics :—

Side stability (in absence of signal current) 4 grammes. Minimum excitation 2.5 ampere turns R.M.S. at 50 c/s. Working excitation for robust operation 4 ampere turns* R.M.S. at 50 c/s.

Contact Pressure at 4 ampere turns R.M.S. 12 grammes. Maximum speed without bias disturbance on contact chatter, 60 c/s. At about 65 c/s. there is some bias, of the order of 5% but above this the Relay works faultlessly up to about 120c/s.

TYPE 5C. As Type 5A but with soldering tags instead of plug-in base.

TYPE 5B. Centre stable model, plug-in base 12 pins. With total contact gap of .002 ins. (i.e. .001 in. on either side) this model has the following characteristics :—

Centre stability 3 grammes.

Minimum excitation 2.5 ampere turns D.C.

Minimum excitation 3.0 ampere turns R.M.S at 50 c/s.

Working excitation for robust operation 4 ampere turns* R.M.S. at 50 c/s.

TYPE 5D. As Type 5B but with soldering tags instead of plug-in base.

A number of standard windings are available up to a maximum of 19,000 turns. A maximum of four separate coils is possible.

* 1 ampere turns corresponds to about 20/Vs. at 50 c/s or 0.7 mW D.C.

Complete details available on request. Manufactured by the Sole Licensees :—

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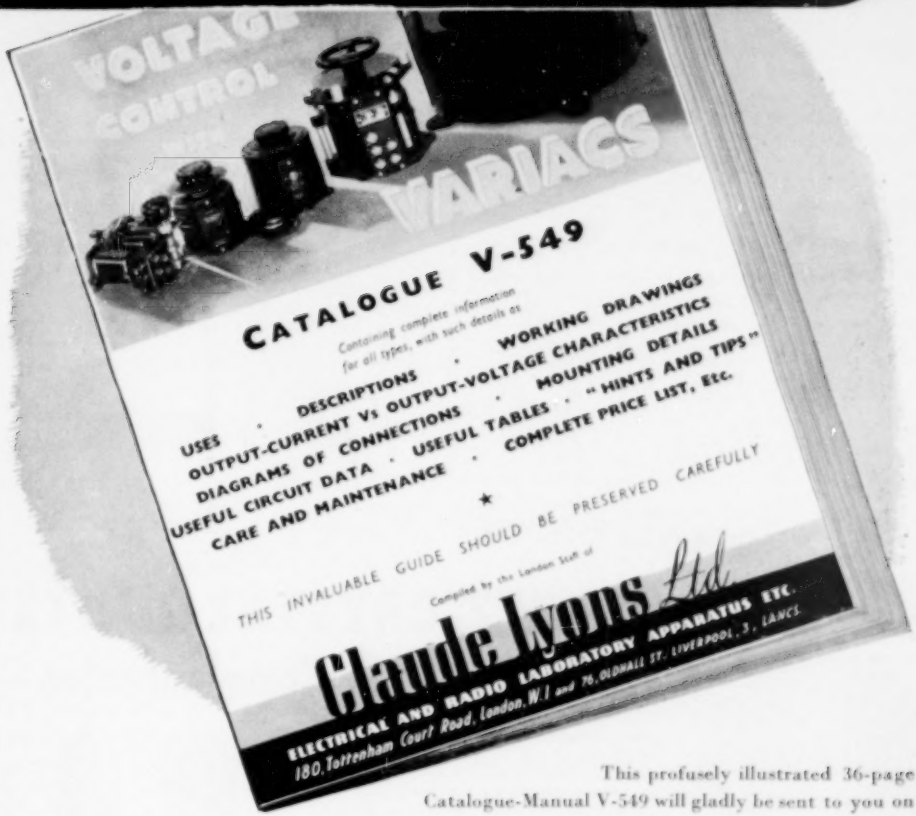
SEND FOR THIS 36-PAGE CATALOGUE


VOLTAGE CONTROL
VARIACS
CATALOGUE V-549
Containing complete information for all types, with such details as
USES · DESCRIPTIONS · WORKING DRAWINGS
OUTPUT-CURRENT Vs OUTPUT-VOLTAGE CHARACTERISTICS
DIAGRAMS OF CONNECTIONS · MOUNTING DETAILS
USEFUL CIRCUIT DATA · USEFUL TABLES · "HINTS AND TIPS"
CARE AND MAINTENANCE · COMPLETE PRICE LIST, Etc.
THIS INVALUABLE GUIDE SHOULD BE PRESERVED CAREFULLY
Compiled by the London Staff of
Claude Lyons Ltd.
ELECTRICAL AND RADIO LABORATORY APPARATUS ETC.
180, Tottenham Court Road, London, W.1 and 76, OLDHALL ST. LIVERPOOL, 3, LANCS.

This profusely illustrated 36-page Catalogue-Manual V-549 will gladly be sent to you on request. It gives complete data on our "Variac" Voltage Regulating Transformers together with dimensioned drawings of all types, and much useful information including many circuits and special applications and tables, etc. There is no charge for this invaluable work of reference—all applications should be addressed to :—

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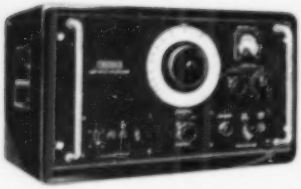

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BRIEF SPECIFICATION

FREQUENCY: 20 c/s—20 kc/s.
ACCURACY: ± 1% or better.
OUTPUT LEVEL: 1W or 10mW into 600 ohms, selected by means of a switch.
ATTENUATORS: 7 steps of 10 db plus 12 db continuously variable.
HARMONICS: 1·75% total at 1 W above 40 c/s 1·0% total at 10 mW above 40 c/s.
HUM LEVEL: 50 db below output level.
DIMENSIONS: 20½-in. x 12½-in. x 11-in. bench mounting — can be supplied for 19-in rack mounting if required.

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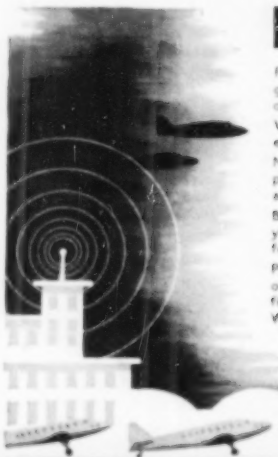
PRECISION MUIRHEAD ELECTRICAL INSTRUMENTS

MU. 24.

G.E.C. QUARTZ CRYSTAL UNITS

For Reliable Radio Communications



From take-off to landing, the safety of passengers, cargo and machines is largely dependent on trouble-free radio communication.

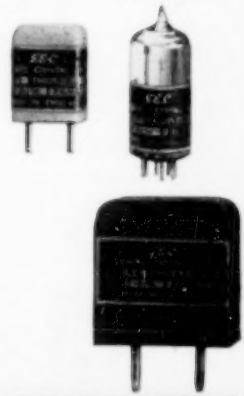
Weather reports, conditions at terminal airports etc., must be reported to aircrews.

Navigation by radio beacons plays an important part in keeping the planes flying safely on their assigned routes.

Be sure, then, that the Quartz Crystal Units in your radio equipment are the best obtainable, for they are the most critical components.

Pioneering research and experience in making over a million units, ensure a product of the finest available quality at competitive prices. We can give rapid delivery for your urgent requirements.

When the requirement is exceptionally urgent, phone Middleton (Lancs.) 2424 or Temple Bar 4669 (London).



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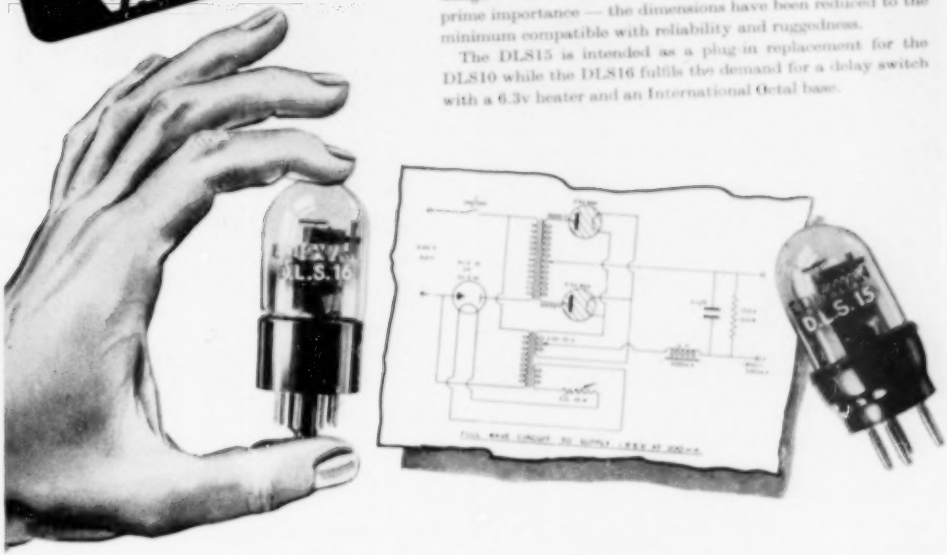
A Subsidiary of THE GENERAL ELECTRIC CO. LTD. OF ENGLAND

EDISWAN
SPECIAL PURPOSE
VALVES

Smaller Vacuum Thermal Delay Switches

These two new delay switches are recommended for use wherever a switching delay of 30-180 secs. is required. They have been developed for use in applications similar to those of the EDISWAN DLS10 but, in line with modern equipment design — which makes the physical size of the switches of prime importance — the dimensions have been reduced to the minimum compatible with reliability and ruggedness.

The DLS15 is intended as a plug-in replacement for the DLS10 while the DLS16 fulfils the demand for a delay switch with a 6.3v heater and an International Octal base.



DLS15 Rating.

Filament Voltage (volts)	4
Filament Current (amps)	0.75*
Delay time at 4 volts (secs.) ...	Min. 30 ; Max. 90

(Delay time may be increased to 180 secs. by variation of filament voltage)

Maximum Peak Current (low voltage rating) 5 amps at 240v.

Maximum Peak Current (high voltage rating) 100mA at 1 kv.

Maximum overall length (not including pins) 60 mm.

**at approximately 4 volts*

DLS16 Rating.

Filament Voltage (volts)	6.3
Filament Current (amps)	0.48*
Delay time at 6.3 volts (secs.) ...	Min. 30 ; Max. 90

(Delay time may be increased to 180 secs. by variation of filament voltage)

Maximum Peak Current (low voltage rating) 5 amps at 240v.

Maximum Peak Current (high voltage rating) 100mA at 1 kv.

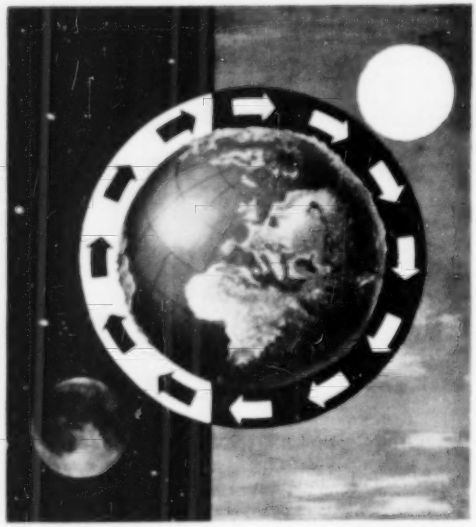
Maximum overall length (not including pins) 65 mm.

**at approximately 6.3 volts*

EDISWAN

THE EDISON SWAN ELECTRIC CO. LTD., 155 CHARING CROSS ROAD, LONDON, W.C.2

R.S.195



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Noon over Nanking. Dawn over the Atlantic. Blackness over the Andes. In all parts of the world, at all times of the day and night, Marconi built stations are in operation, broadcasting or sending telegraph messages and giving unrivalled service. Wherever ships and aircraft are plying, you will find Marconi apparatus guiding them, every hour of the day, every day of the week. Pioneers in Wireless Communication fifty years ago, Marconi's still maintain their leadership. Behind every new development there will be Marconi's accumulated experience. Behind every new piece of Marconi equipment there will be the honoured name as an assurance of maximum efficiency.

Marconi

the greatest name in wireless

MARCONI'S WIRELESS TELEGRAPH COMPANY LTD.,

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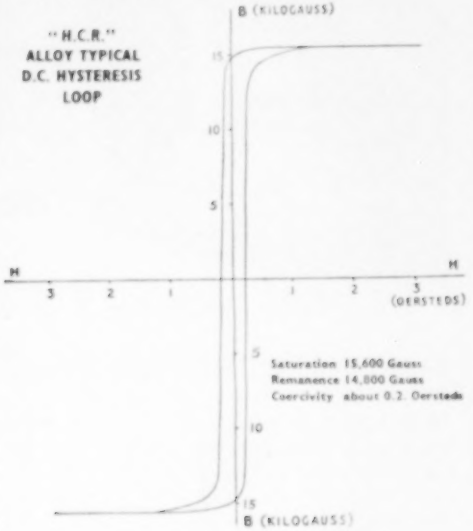
TELCON'S LATEST MAGNETIC DEVELOPMENT

H.C.R. ALLOY

is now available in commercial quantities



The magnetic properties of this alloy make it the ideal core material for

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NITROGOL *Capacitors*

QUALITY has always been the basis of Dubilier technical supremacy and in the television field too, this factor has confirmed Dubilier's lead in the international field.

Thus it is natural that only components of tested quality and worth will satisfy television engineers. Experience has proved that Dubilier Nitrogol Capacitors, acknowledged by technicians throughout the world for their reliability, are best suited to the manufacturer of television apparatus.

In such applications as radar, amplifiers, undersea cable repeaters and industrial electronic applications, Nitrogol Capacitors are again a prime necessity

Comprehensive catalogues giving detailed information of capacitors and technical service are available upon request.

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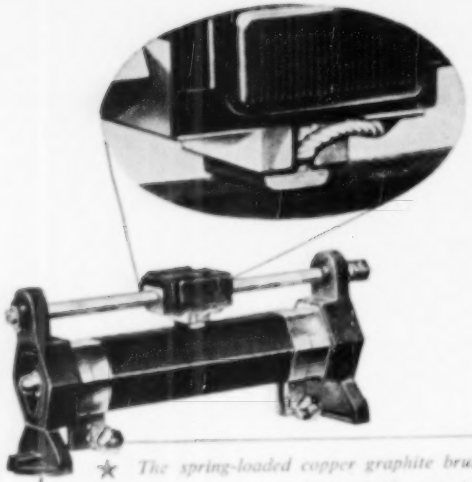
'Phone: Acorn 2241 (5 lines). 'Grams: Hivoltcon, Wesphone, London. Cables: Hivoltcon, London. Marconi International Code

D38A



PERFECT CONTACT★

To ensure perfect contact at all temperatures and to prevent undue wear of the windings BERCO sliding rheostats and potentiometers are fitted with a spring-loaded copper graphite self-lubricating brush operating on the flat surface of a hexagonal solid drawn steel tube.

Open, protected or ganged types are available in a wide variety of sizes. Graded windings can be supplied for special applications.

Ask for leaflet No. BR601/115

★ *The spring-loaded copper graphite brush is held accurately in alignment in a diecast holder, providing a permanently lubricated contact at high temperature. The pigtail connection ensures current is not carried by the springs.*

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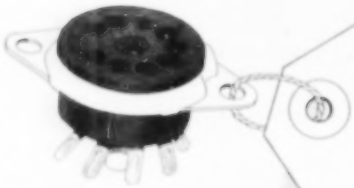
SLIDING RESISTANCES

THE BRITISH ELECTRIC RESISTANCE CO. LTD.
QUEENSWAY, PONDERS END, MIDDLESEX. *Phone:* Howard 1492. *Grams:* Vitrohm Enfield.

Recognised as the Most Reliable Valveholders



McMURDO
Moulded
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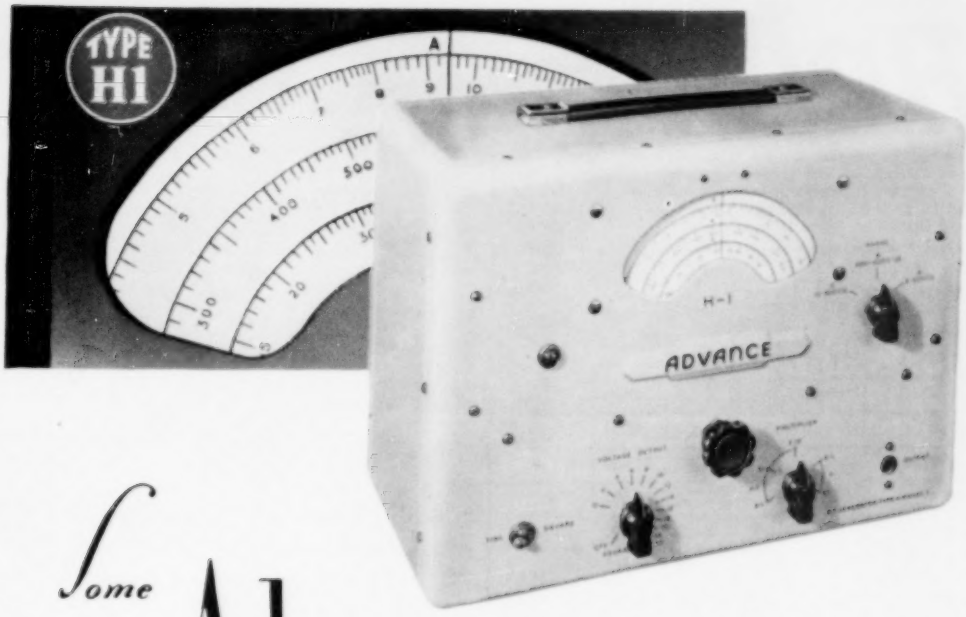
B9A—*NOVAL*

Ref. No. **BM9/U**

Moulded in **Phenol-Formaldehyde**

* *Also supplied with various saddle assemblies to meet customers' requirements.*

THE McMURDO INSTRUMENT COMPANY LTD., VICTORIA WORKS, ASHTEAD, SURREY · ASHTEAD 3401



Some **Advance** features in

AUDIO GENERATOR design

THE combination of wide-range, high accuracy and low distortion will commend the Advance H.1 Audio Generator to all critical Service Engineers. The 15 c/s to 50 Kc/s coverage is achieved by the use of a bridge type resistance-capacity oscillator with three switched ranges. Stability throughout is attained by a highly efficient stabilizing circuit followed by two amplifier stages with heavy negative feedback.

Output : Sine wave 200 microVolts to 20 Volts, plus/minus 1 db, or Square wave 400 microVolts to 40 Volts (800 microVolts to 80 Volts peak to peak). Measured distortion is less than 1% at 1,000 c/s. In attractive cream enamel sprayed steel case, 13¼" x 10¼" x 8". Weight only 14 lbs.

LIST PRICE **£20**

**The Scale reproduced above is actual size.*

COVERS
15 c/s to 50 Kc/s
in three stages

ACCURACY
plus-minus 1%
plus-minus ..... 1 c/s

SINE & SQUARE WAVE OUTPUT

MAX. POWER OUTPUT— UNDISTORTED
0.25w into 2,000 ohms.

ADVANCE COMPONENTS LTD
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ROLA

A name which ten million speakers have carried to the uttermost parts of the earth. A range which provides a unit to meet all normal requirements. Brief details are given below.

Model	Overall Diameter	Magnet Pole Diameter	Flux Density	Total Gap Flux (Maxwell)	Peak Power Capacity
B3KO	3" (sq.)	.57"	7,500	14,000	.5 w.
B4KO	4" (sq.)	.57"	7,500	14,000	.75 w.
B5K	5"	.57"	7,500	14,000	2 w.
K5Q	5"	.75"	8,500	21,000	2.5 w.
K5G	5"	.75"	10,000	25,000	3 w.
K6D	6.5"	.75"	6,500	16,500	2.75 w.
K6Q	6.5"	.75"	8,500	22,000	3 w.
K6G	6.5"	.75"	10,000	26,000	3.5 w.
K46D	6.25 x 4.25 (Oval)	.75"	6,500	16,500	2.5 w.
K46Q	6.25 x 4.25 (Oval)	.75"	8,500	22,000	3 w.
K46G	6.25 x 4.25 (Oval)	.75"	10,000	26,000	3 w.
F6JO	6.5"	.6"	9,000	18,500	3 w.
Z8D	8"	1"	6,500	25,000	4 w.
Z8M	8"	1"	8,000	31,000	4.5 w.
Z8R	8"	1"	9,500	37,000	5 w.
Z10D	10"	1"	6,500	25,000	6 w.
Z10M	10"	1"	8,000	31,000	6.5 w.
Z10R	10"	1"	9,500	37,000	7 w.
P44	12.2"	1.5"	10,000	60,000	10 w.
G.12	12.2"	1.75"	13,000	146,000	12 w.

Voice Coil Impedance is 3 ohms in the case of all the above models except the G.12, where the impedance is either 15 or 8 ohms.

Write for details—enquiries are invited for Cones, Laminations and Output Transformers.

BRITISH ROLA

Ferry Works, Summer Rd., Thames Ditton, Surrey.

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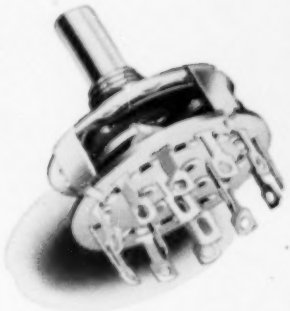
THE SPEAKER YOU KNOW BY EAR

Which switch fits the niche?



By the time that has rolled round your tongue half a dozen times the Type B.T. Switch could be fitted! It's that sort of switch—small, compact, designed specially to fit easily into small spaces. It'll save you a lot of trouble. We have recently included some new features. Here they are—

- **The contact blades are fixed in a new way. This makes them completely rigid all the time.**
- **The drive spindle is positioned definitely in the wafer. This provides double bearing.**
- **Steady contact resistance.**
- **Improved insulation between contacts.**
- **Positive contact.**
- **Self-cleaning action.**

All these make the Type B.T. a better switch—more efficient and lasting. In fact just the switch you need for limited space and simple switching.

WALTER SWITCHES—MADE FOR LIFE

Walter Instruments

LIMITED

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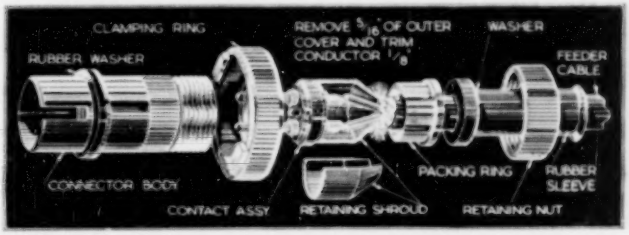
CRC47

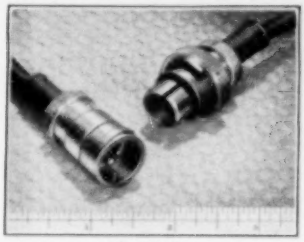
ONE OF THE NEW LINES AT THE R. E. C. M. F. EXHIBITION 1950

(STAND No. 66)

'SCREENECTORS' A NEW RANGE OF SCREENED CONNECTORS

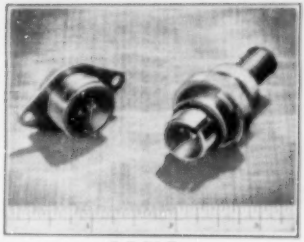


SHOWING METHOD OF LOADING (ACTUAL SIZE)



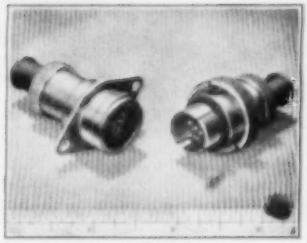
COAXIAL

L.722 P	Flex plug. Above right.
L.722 S	Chassis socket not shown.
L.723	Through Chassis socket not shown.
L.724	Flex socket. Extreme left.



2-POLE

L.625 P	Flex plug. Above right.
L.625 S	Chassis socket. Above left.
L.689	Through Chassis socket not shown.
L.690	Flex socket not shown.



3-POLE

L.715 P	Flex plug. Above right.
L.715 S	Chassis socket not shown.
L.716	Through Chassis socket. Above left.
L.717	Flex socket not shown.

'SCREENECTORS'

This new range of screened connectors in light alloy is based upon the draft R.E.C.M.F. specification for a non-reversible screened plug and socket to load cables up to 0.24 in diameter over the braid.

Single or multipole contacts may be assembled into the common housings which are designed so that various applications may be employed, *e.g.* line connector, flex plug and chassis socket and *vice versa*, through chassis (bulkhead) connector, each part interchangeable in its appropriate position in the assembly.

Other design features include: **(1) High grade bakelite insulators Nylon X17163 or equivalent. (2) Simple assembly and loading. (3) Positive quick action locking device. (4) Machined light alloy screened housing finished for instrument panel requirements.**

The characteristics of these connectors are shown opposite.

Type	Characteristic impedance ohms★	Contact resistance	Capacitance★	
			Conductor/ conductor	Conductor/ screen
Coaxial	75	Less than 5 milliohms each	—	—
2-pole	100		1.0 pF	2.5 pF
3-pole	—		—	—

★ at 1 Mc/s

OTHER NEW LINES

In addition to the "Screenectors" described here we hope to show a reliable 10mA fuse ("Minifuse" Pat. app. for); valveholders for B7G, B9A and for Duodecal C.R.T. base.

Aerials will not be forgotten, and the following television aerials introduced since the 1949 show will be exhibited: **"Multirod"** 4 element array for fringe areas; **"Viewrod"** 'H' type 0.15 wave spacing; **"Twinrod"** outdoor dipole, windowsill or gutter board mounting dual purpose aerial. **"Viewflex"** all flexible indoor dipole. The popular **"Doorod"** indoor aerial will be shown together with the well known **"Skyrod"**, **"Eliminoise"**, **"Winrod"**, and **"Carod"** broadcast aerials.

Our representative will welcome all enquiries at our **Stand No. 66.**

BELLING & LEE LTD

CAMBRIDGE ARTERIAL RD., ENFIELD, MIDDX., ENGLAND

DEVIATION TEST BRIDGE

Type No. "S" 1502 and 1507

indicates on a large illuminated scale the percentage deviation from an arbitrary impedance. The bridge is particularly suitable for checking resistors, capacitors and inductors in industrial plants, but has also proved useful in many special measurements based on comparison between two impedances.

Specification :

Measuring Objects for Deviation Measurements : Resistance 10Ω–10 MΩ capacitance 50 pF–10 μF, inductance 2 mH –100 H for type 1502. Capacitance 25 pF–0.1 μF, inductance 100 μH–0.3 H, resistance 20Ω–100kΩ for type 1507.
Measuring Objects for Direct Reading with External Standards : Resistance ranges 0–0.7Ω to 0–3.3 MΩ, capacitance ranges 0–7 pF to 0–3.3 μF for type 1502 0–7 pF to 0–0.033 μF for type 1507
Measuring Frequency : 1000 c/s for type 1502 and 50 kc/s for type 1507.
Instrument : Large moving-coil instrument permits easy readings. Illuminated scale with zero point in the centre. The instrument is safeguarded against overloads. Full scale deflection at 7% and 25%.
Comparison Accuracy : Better than 0.1% at zero point. The zero point is independent on the line voltage.
Tolerance Accuracy : Better than 5% at full scale deflection.
A.C. Hum : The bridge is not sensitive to hum voltages.
Power Supply : 115–127–150–220 volts a.c. Power consumption 40 watts.
Dimensions : 13 (height) × 11 (width) × 8 (depth excl. dials) inches.

Manufactured by

Brüel & Kjær

Nærum, Denmark

NEW

SERIES 800 ELLIPTICAL (Type 874)

CAN 4–8? Of course it can't. And a 4" speaker can no more equal the performance of an 8" than a quart can come out of a pint pot.

BUT that is no reason why a 4" speaker cannot be a really good 4" speaker. Precision in machining, exactness of assembly and, of course, CO-AXIAL ALIGNMENT of all component parts is just as important as in any other of our speakers.

Dimensions and Data

Horizontal	7.1" (18 cm.)
Vertical	4.1" (10.3 cm.)
Axial	2.4" (6 cm.)
Flux Density	8,000 or 9,000 Gauss

We are really pleased with our 4" × 7" Elliptical. So will you be when you hear it.

REPRODUCERS AND AMPLIFIERS LIMITED
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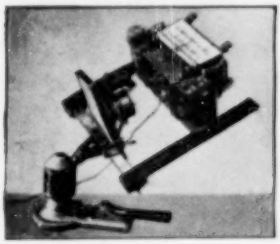

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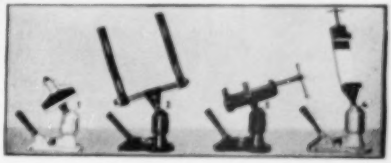

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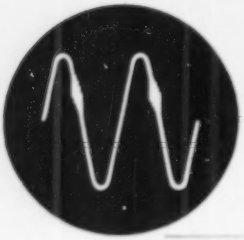

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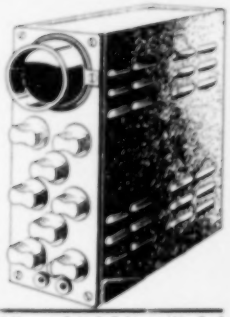

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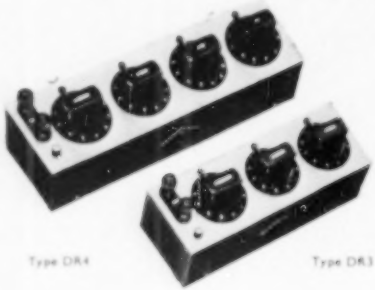

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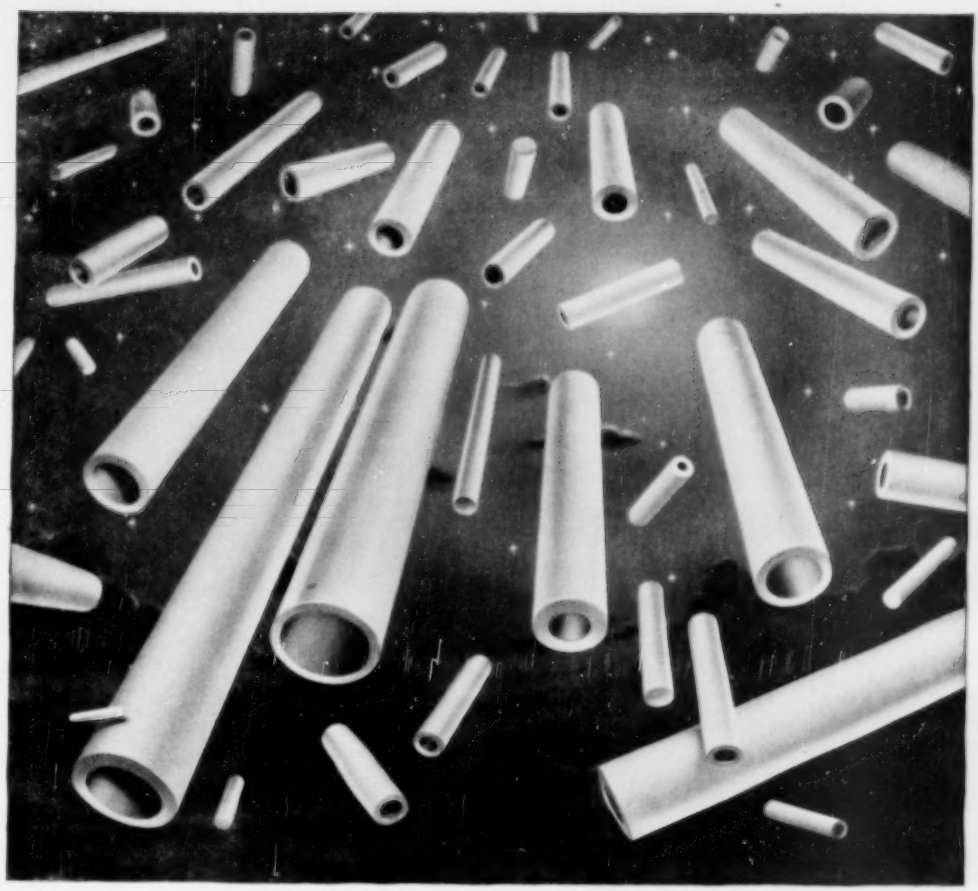

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APRIL 1950

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Type	f (Mc/s)	Va (V)	Vg2 (V)	Vg1 (V)	Ia (mA)	Ig2 (mA)	Ig1 (mA)	Vin (pk) (V)	Pout (W)	η (%)	Pload approx (W)
TY2-125	150	2500	—	−200	200	—	40	350	360	72	290
TY3-250	70	3000	—	−250	373	—	70	420	850	76	680
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EA76	Single Diode. (5mm bulb)
EC70	R.F. Triode for use as Oscillator up to 500 Mc/s
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WIRELESS ENGINEER

Vol. XXVII APRIL 1950 No. 319

Can *H* and *D* be Measured Directly?

IN the Editorial of August 1946 we discussed the analogy between mechanical stress and strain, electrical $\mathscr{E}$ and D and magnetic H and B. This is a question that is continually arising, and on which divergent views are held. Strain can be measured directly but in his 'Theory of Elasticity' Professor Southwell expresses the opinion that stress has never been measured directly and "we can assert with some confidence that it never will." The stress can, however, be determined, at least theoretically, by indirect measurement, such as measurement of the strain produced in a standard material inserted at the point under consideration, analogous to the crevasse cavity experiment in magnetism. The stress is, however, usually calculated by dividing the load force by the cross-sectional area. In the Editorial referred to above we maintained that a similar state of affairs existed in the electric and magnetic fields; that in each case we have a directly measurable magnitude, B or $\mathscr{E}$ and a magnitude that cannot be directly measured, H or D. We showed that when a magnetic needle or solenoid is used to determine H, the measurement is primarily of the magnetic induction B in which it is situated, and that H is determined indirectly by what is really a cavity experiment. H can be calculated by dividing a current or a number of ampere-turns by a length and similarly D can be calculated by dividing a charge by an area.

This point of view has recently been discussed by P. Cornelius and H. C. Hamaker in *Philips Research Report* No. 4 of April 1949, and they maintain that both H and D can be directly measured. It depends largely on what is meant by the direct measurement of a magnitude which is really a condition existing at a given point. In a uniformly-wound toroid one can measure the current with an ammeter, and the length of the magnetic path in the toroid with a cm-scale, and divide one by the other to obtain the ampere-turns per cm. Can one call this a direct measurement of H? Cornelius and Hamaker say that H may be measured, at least theoretically by means of a small solenoid of a single turn constructed as shown in Fig. 1, its diameter being small compared with its length. The material is assumed to be perfectly conducting and the coil is assumed to be short-circuited through an ammeter of negligible internal resistance. If the coil is brought from a field-free place to the point where H is to be measured a permanent current I will be generated and H in the direction of the axis is defined as the limit of I/l as the length l of the coil approaches zero about the point at which H is to be determined. All this is very theoretical—one might almost say fictional—and tends to confirm the view that H is not directly measurable. Moreover it is important to notice that the current set up in the short-circuited ideal solenoid produces a value of H sufficient to maintain the flux through the solenoid, and therefore also B, at zero. This is not necessarily the same as maintaining the resultant H at zero, unless the medium is free from hysteresis. If the medium had residual magnetism with $H = 0$, then on

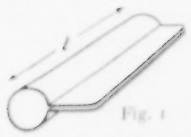

Fig. 1

introducing the short-circuited solenoid a current would be set up which would be a measure of the coercive force and not of H.

With the measurement of the displacement D the authors seem to be on somewhat safer ground. They state that the method was indicated in Pohl's "Einführung in die Electricitätslehre." Assuming the medium to be gaseous or liquid, two metal discs are brought into contact face-to-face at the point where D is to be measured, then separated and removed to a field-free place where they can be discharged through a ballistic galvanometer and the charge Q thus determined. D at the point will be equal to the quotient Q/A as the area A of the discs approaches zero. Here again, however, the magnitude D at the point is not measured, but is assumed to be calculated from the limiting value of the quotient of a quantity and an area as the area approaches zero,—a very fictitious experiment. When two small metal discs are placed in contact face-to-face in the dielectric field, charges accumulate and distribute themselves over the external surfaces so as to make the resultant electric field $\mathfrak{E}$ in the metal zero at every point. It might be thought at first sight that this charge would depend on the value of $\mathfrak{E}$ and not of D in the dielectric. It must be noted, however, that this is a cavity experiment; the metal discs are occupying a crevasse in the dielectric, and in the absence of the metal, the value of $\mathfrak{E}$ in the vacuous crevasse would be κ times that in the dielectric, since the displacement D would be continuous across the crevasse. Hence the charge on the discs is a measure of $\kappa\mathfrak{E}$, that is, of D and not of $\mathfrak{E}$ in the dielectric.

Professor Southwell did not say that the mechanical stress at a point could not be determined; he said that it could not be measured directly, and the arguments of Cornelius and Hamaker seem to confirm the view that the same is true of H and D.

G. W. O. H.

The Movement of Electrons in a Transmitting Aerial

IT is now generally known that the old idea that the current in a conductor is due to a rapid movement of electrons is quite wrong, and that even very large currents are associated with extremely slow drift of the electrons of the order of a few millimetres per second. If an alternating current of 1,000 amperes at a frequency of 50 cycles per second flows in a conductor of 1 square centimetre cross-section, the mean value of the current is about 900 amperes, and the total quantity crossing any cross-section in a half-cycle (i.e., in 1/100 second) is 9 coulombs. Assuming the number of free electrons to be equal to the number of atoms (i.e., 84×10^{21} per cm^3) and the charge on each to be 1.6×10^{-19} coulomb, the quantity of free electricity per cm^3 is $84 \times 1.6 \times 100$; i.e., 13 440 coulombs. In order that 9 coulombs may pass across the 1 cm^2 cross-section in the half-cycle, the average movement or drift of the electrons must be 9/13 440 cm. Hence we see that in a conductor carrying an alternating current of 1,000 amperes at 50 cycles per second the extreme travel of the free electrons does not exceed about a thousandth of a centimetre.

The results are even more surprising when we consider a radio transmitting aerial, for we feel sure that most people picture something rushing up and down the aerial. If the current is 100 amperes and the wavelength 10 metres, then assuming the aerial to have a cross-section of 0.5 cm^2, the average current density will be 180 A/cm^2, and the time of a half-cycle $1/(60 \times 10^6)$ second. The total charge of the free electrons in 1 cm length of the aerial will be 6 720 coulombs, and if their movement in the half-cycle is equal to x, then $6\,720 \times x = i_{av} \times t = 90 \times 1/(60 \times 10^6)$ from which we find that the total movement x of the electron is equal to $2.2/10^{10}$ cm. Far from rushing up and down the aerial, the free electrons—and they are the only things that move—have only a sub-microscopic movement.

During the quarter-cycle that the aerial is charging the average current is 90A and the duration of the quarter-cycle $1/(120 \times 10^6)$; hence the charge imparted to the aerial will be $75/10^8$ coulomb, and if the capacitance is say 100 pF, the aerial will be charged to a voltage of $(75/10^8) \times 10^{10} = 7{,}500$ volts. Hence, although the movement of the free electrons is sub-microscopic, their density is so great that the movement produces an appreciable change in the distribution of the charge and the consequent potential distribution.

The same reasoning would apply to the classical experiments of Hertz with his oscillator and spark-gap. Although the electron was then unknown, we feel sure that everybody at that time pictured the breaking down of the spark-gap as initiating a rushing to and fro of electricity in some form or other. This was doubtless true in the spark-gap, but in the adjacent conductors the movement of the electrons would be almost infinitesimally small.

G. W. O. H.

HIGH-STABILITY OSCILLATOR

By G. G. Gouriet, A.M.I.E.E.

(*Research Department, Engineering Division, B.B.C.*)

SUMMARY.—An oscillator circuit which has been used extensively by the B.B.C. for some considerable time is discussed and analysed.

It is shown that with the circuit designed for maximum frequency stability, the stability obtained is for all practical purposes, a function only of the parameters of a single tuned circuit at series resonance.

Measured performance data are given for a practical form of the circuit which has been used for many years as a variable-frequency source of drive for B.B.C. short-wave transmitters.

1. Introduction

DURING the past year several references have been made in the American technical press to a particular form of oscillator circuit which bears a close resemblance to a Colpitts oscillator, the main difference being that a series LC circuit is substituted for the tuning inductance of the Colpitts circuit[1,2]. The circuit of this oscillator was conceived and developed independently in this country by the author some nine years ago, initially for use as a crystal-controlled oscillator, and in the same year as an LC controlled oscillator for use in variable-frequency transmitter-drive equipment.

Since its development, this circuit has proved to be highly satisfactory and has been used almost exclusively in B.B.C. transmitter equipment throughout the country.

Some aspects of the th oretical behaviour of this oscillator have been dealt with by J. K. Clapp, in an article published early last year.[1] It is, however, proposed to present here the author's original analysis of the circuit, since this is considered to be somewhat simpler than the treatment by Clapp, while it nevertheless demonstrates the many unique features of the circuit and provides the necessary design information for obtaining optimum performance.

Before discussing the circuit in question, it is considered worth while to examine briefly some of the fundamental principles which apply in general to all oscillators.

Any network that is capable of sustaining continuous oscillation at a fixed frequency must fundamentally comprise a frequency-controlling element and a source of energy. In order that a condition of stable equilibrium may be satisfied, any practical oscillator must also contain some non-linear element which is capable of ensuring that the amount of energy supplied to the frequency-controlling element will compensate exactly for the energy expended in the latter due to inherent losses. However, in theory at least, we are entitled to simplify the analysis by making the assumption that the equilibrium is obtained by an initial adjustment to the source of energy and thereafter that no changes occur in the value of the parameters involved, and for this idealized case we may regard all elements as being linear, and base our analysis on the assumption of a perfectly sinusoidal oscillation. While such an ideal arrangement cannot be realized in practice, we shall show later that it can be so closely approximated as to make all the theoretical conclusions substantially correct for the case of a practical circuit.

Neglecting the possibility of compensating errors, it is quite obvious that the frequency stability of an oscillator can never exceed that of its frequency-controlling element; for example, the greatest accuracy with which a clock can record time is strictly limited to a figure which is dependent upon the physical stability of its pendulum. The accuracy obtained from the clock will probably be substantially less than this limiting accuracy, the reason being that it is impossible to impart to the pendulum the energy that is necessary to overcome the frictional losses, without altering to some extent its period of free oscillation. Thus, when comparing the merits of different oscillator circuits, it is convenient to assume in each case that the frequency-controlling element is intrinsically stable, and to examine the extent to which the 'pendulum' is upset by the presence of the maintaining element.

Fig. 1. *Basic series-resonance circuit.*

The simplest electrical frequency-controlling element is the tuned circuit comprising an inductor and a capacitor. As is well known, with such a circuit the loss of energy occurs mainly in the inductive limb, and Fig. 1 is a close approximation to a physically realizable closed resonant circuit.

In order to replace the energy dissipated in the loss resistance R, the simplest procedure, were it possible, would be to insert into the loop a two-terminal impedance of such a value that the circulating current i would develop across it a

MS accepted by the Editor, September 1949.

voltage equal to that developed across the loss resistance, but of opposite sign as shown in Fig. 2(a). The value of this impedance would, of course, be $-R$, which introduces the concept of negative resistance. It is apparent that such an element would be providing energy and not dissipating it, and furthermore it would be a perfect maintaining element since it would have no effect upon the loop resonance frequency at which the oscillation would be maintained.

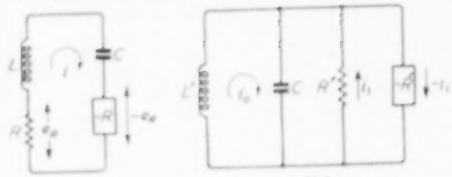

Fig. 2. *Zero-loss circuit obtained with series negative resistance (a), or shunt negative resistance (b).*

There is an alternative method by which energy could be supplied to the circuit. As is well known, we may express the series loss resistance as an equivalent parallel resistance R' and arrive at the circuit of Fig. 2(b). The maintaining element may, therefore, be connected across the inductive limb, in which case, if expressed as a two-terminal impedance, it would need to have a value $-R'$.

It will be seen that there are two distinct disadvantages in this arrangement; first, the frequency of oscillation is no longer independent of the value of R, and secondly, the impedance across which the maintaining element is connected is relatively high and is, therefore, more susceptible to the effect of stray shunt capacitance. While it is true that the arrangement of Fig. 2(a) is equally susceptible to stray series inductance, the latter is far less troublesome at the frequencies for which such oscillators are normally used.

The ideal arrangement of Fig. 2(a) cannot be realized in practice, but circuits which will provide a

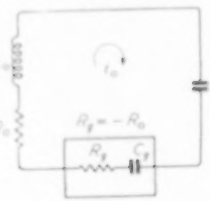

Fig. 3. *The impedance of the maintaining circuit is represented by $R_g C_g$.*

complex impedance that includes a negative resistance are quite common and in fact form part of any oscillator circuit. Such a case is shown in Fig. 3, where the impedance of the maintaining circuit is represented by a negative resistor R_g and a series capacitor C_g. It is clear that if C_g is small and comparable with C_0, its effect upon frequency will be of a first order, and it must, therefore, be regarded as part of the frequency-controlling element. If it exists physically as a capacitor it may be possible to arrange for it to have the same order of stability as that of the controlling capacitance C_0, but if it merely represents a quadrature component of voltage or current, as, for example, in the case of Miller effect, then to stabilize its value may prove to be very difficult.

In either case, if C_g is sufficiently large compared with C_0, although strictly speaking it will still form part of the frequency-controlling network, its effect upon frequency may be small enough for it to be neglected. In the analysis which is to follow, it will be shown that, in the circuit with which we are concerned, C_g is a capacitance which exists physically, and also that it can be made sufficiently large compared with C_0 for the circuit to approximate closely to the ideal arrangement of Fig. 2(a).

2. Theoretical Analysis

A simplified form of the B.B.C. oscillator circuit is shown in Fig. 4(a). In practice a choke of suitable inductance would be connected across C_2 to serve as a d.c. path for the anode current of the valve, but since this component has virtually no effect on the behaviour of the circuit it has been omitted. If an inductance is substituted for the series LC element which is connected between terminals G and E, the circuit will become identical to a Colpitts oscillator.

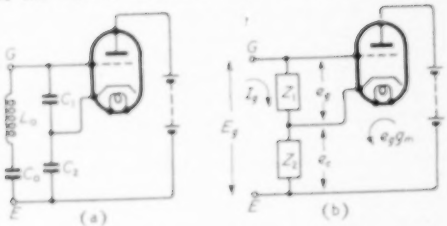

Fig. 4. *Simplified form of the practical circuit (a), and the generalized form of its oscillation maintaining elements (b).*

In Fig. 4(b) the series LC element has been removed, and the capacitors C_1 and C_2 have been replaced by the generalized impedances Z_1 and Z_2. It is shown in Appendix I that the impedance looking into the terminals G and E may be written

$$Z_g = Z_1 + Z_2 + g_m Z_1 Z_2 \quad .. \quad .. \quad (1)$$

From this it will be seen that if Z_1 and Z_2 are pure reactances, jX_1 and jX_2 respectively, the impedance becomes

$$Z_g = j(X_1 + X_2) - g_m X_1 X_2 \quad .. \quad (2)$$

of which the real part will be negative providing X_1 and X_2 are of the same sign. The analysis is therefore equally applicable to a Hartley circuit, similarly modified, in which Z_1 and Z_2 are inductances.

Returning to the circuit of Fig. 4(a) in which Z_1 and Z_2 are the reactances of C_1 and C_2 respectively, we shall obtain here, for the input impedance, a resistance,

$$R_g = -\frac{g_m}{\omega^2 C_1 C_2} \quad .. \quad .. \quad .. \quad (3)$$

in series with a capacitative reactance,

$$\frac{1}{\omega C_g} \text{ where } C_g = \frac{C_1 C_2}{C_1 + C_2} \quad .. \quad .. \quad (4)$$

It will be noted that C_g is simply the series combination of the physically real capacitances C_1 and C_2.

With the series LC element connected across this impedance, it is apparent that the condition of loop resonance which will determine the frequency of oscillation is given by

$$\omega L_0 - 1/\omega C_0 - 1/\omega C_g = 0 \quad .. \quad .. \quad (5)$$

The energy equilibrium necessary for continuous oscillation will be satisfied by the condition

$$R_0 = -R_g = g_m/\omega^2 C_1 C_2 \quad .. \quad .. \quad (6)$$

By letting G equal the maximum value of mutual conductance that can be obtained from the valve under prescribed conditions, we may allow for slightly more energy to be available than is actually required to maintain the oscillation, and write (6) as

$$R_0 \leqslant G/\omega^2 C_1 C_2 \quad .. \quad .. \quad .. \quad (7)$$

Now it is desirable that the total capacitance C_g shall be as large as possible in order that the frequency of oscillation shall approximate closely to the resonant frequency of the controlling elements L_0 and C_0, but a limit to the maximum value is imposed by (7), which states implicitly that for a given value of mutual conductance and for a given frequency, unless the product $C_1 C_2$ is sufficiently small to satisfy the equilibrium, the energy supplied will be insufficient to maintain the oscillation.

For a given product of C_1 and C_2, the series capacitance C_g is a maximum when $C_1 = C_2$, and by writing

$$C_1 = C_2 = 2 C_g \quad .. \quad .. \quad .. \quad (8)$$

we obtain from (7)

$$1/\omega C_g \geqslant 2\sqrt{(R_0/G)} \quad .. \quad .. \quad .. \quad (9)$$

This equation is important since it shows that, for oscillation to be maintained, the minimum permissible reactance of C_g is a function only of the loss resistance of the frequency-controlling element and the valve mutual conductance.

At this stage in the analysis it is worth while examining the value that this reactance will have in a practical case. For example, we will assume that a high-slope pentode is used with a maximum mutual conductance of 8 mA/V and that it is operated at a value of 6 mA/V.

At a frequency of 1.0 Mc/s let

$$\omega L_0 = 1/\omega C_0 = 800\ \Omega\ (L \approx 127.5\ \mu\text{H}, C = 200\ \text{pF}) \text{ and } Q = \omega L_0/R_0 = 200$$

whence $R_0 = 4\ \Omega$

The required reactance of C_g, assuming that the oscillation frequency is 1.0 Mc/s, is thus

$$X_g = 2\sqrt{\{4/(6 \times 10^{-3})\}} = 51.6\ \Omega \quad .. \quad (10)$$

which corresponds to a capacitance,

$$C_g = 3100\ \text{pF approximately.}$$

This capacitance is in series with the frequency-controlling capacitance C_0, and therefore the oscillation will occur at a frequency slightly higher than 1.0 Mc/s. To a first approximation the actual frequency will be obtained by multiplying 1.0 Mc/s by a factor $\sqrt{(1 + C_0/C_g)}$.

Since the value of C_0 is approximately 200 pF we obtain for the actual frequency of oscillation, to a close approximation

$$\sqrt{(1 + 200/3100)} = 1.032\ \text{Mc/s} \quad .. \quad (11)$$

The error of 3.2% is sufficiently small to be neglected in the following analysis and henceforth we shall assume that the oscillation occurs at the series resonant frequency of L_0 and C_0.

In the practical case we have just examined, the capacitors C_1 and C_2 will each be approximately 0.006 μF, and at the frequency of oscillation the reactance of each will be of the order of 25 Ω. These figures indicate how relatively insignificant will be the effect upon frequency of stray impedance, due, for example, to the valve interelectrode capacitances, since the magnitude of such impedance at the frequency of oscillation will be of the order of thousands of ohms.

In addition to the presence of stray capacitance, the value of the impedance Z_g will also be modified to some extent owing to the presence of harmonic components which, so far, we have chosen to ignore. Llewellyn has shown[3] that by the process of intermodulation these components can cause an additional phase-shift at the fundamental frequency. The significance of such phase-shift will be discussed in more detail later, but for the present, assuming that a modification to the impedance Z_g results, it is apparent that it can be accounted for completely in terms of a change of C_g, since the real part of Z_g must of necessity exactly equal the loss resistance R_0, which according to our premises is of constant value.

Thus the merit of the circuit, from the point of view of frequency stability, is conveniently expressed by the value of the differential coefficient, df/dC_g and in Appendix 2 this has been

deduced for the general case and is shown to be given by

$$\frac{df}{dC_g} = -\frac{4\pi f^2}{QG} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (12)$$

For small changes of the order of a few parts in a million we may write this as

$$\frac{\Delta f}{f} = -\frac{4\pi f}{QG} \cdot \Delta C_g \quad \ldots \quad \ldots \quad \ldots \quad (13)$$

or $$\Delta F = -\frac{4\pi F}{QG} \text{ cycles/megacycle/per pF} \quad (14)$$

where F is expressed in megacycles per second.

As a generalized expression of the stability, the equation in this form is very useful, since we can obtain from it immediately the minimum value of the coefficient that can be obtained at any frequency, given the Q of the controlling element and the maximum mutual conductance at which the valve can operate.

At first sight, however, this expression is likely to be misleading, since it would suggest that for circuits of similar Q the coefficient is independent of the L/C ratio of the controlling element, and also that its magnitude is directly proportional to the frequency of oscillation. While this is quite correct, it is somewhat surprising on philosophical grounds, but the reason is bound up with the postulate of a constant Q. The time constant of an inductance is physically a more appropriate expression of its merit than is its Q, and if we substitute $\omega L_0/R$ for Q in, equation (13) we obtain $\frac{\Delta f}{f} = -\frac{2}{\tau G} \cdot \Delta C_g$ where

$$\tau = L_0/R_0 \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (15)$$

This is entirely reasonable since it states simply that the stability depends only upon the loss factor of the controlling element and the mutual conductance of the valve, the latter of which parameters expresses in effect the degree of coupling which exists between the maintaining element and the controlling element.

Since in practice the value of R_0 depends upon such factors as skin effect and eddy-current loss, the effective value of τ at any particular frequency may be obtained conveniently only by steady-state measurements, and for this reason the coefficient expressed in terms of Q is more useful. If it is found that over a frequency band the maximum value of Q that we can obtain tends to remain constant, so that the fractional frequency change expressed in (14) increases with frequency, it is simply because we cannot produce such good inductors at high frequencies.

As a practical example, if we substitute in (14) the values quoted in the previous example, in which $F = 1.0$ Mc/s, $Q = 200$, $G = 6$ mA/V we shall obtain $\Delta F = \frac{4\pi}{200 \times 6 \times 10^{-3}} = 10.5$ cycles per second per pF change of C_g. A similar change in the value of C_1 or C_2 alone will produce one quarter of this change in frequency, which is approximately 2.6 cycles per second per pF.

For such small frequency changes to be of significance it would, of course, be necessary to stabilize the temperature of the controlling elements L_0 and C_0 to a relatively high degree of constancy, and if this is done, there is no reason why the capacitors C_1 and C_2 should not be included in the temperature-controlled chamber.

In this connection it is of interest to compare the frequency : temperature coefficient of the input capacitance C_g and the frequency-controlling capacitor C_0. High-stability capacitors are available commercially with a capacitance temperature coefficient as low as 25 parts in 10^6 per °C over a wide range of values. If the values of C_1 and C_2 are each changed by 1.0 pF simultaneously, the resulting change in frequency for the values quoted in the above example will be 5.25 c/s. For the frequency : temperature coefficient we thus have

$$\Delta f = \frac{5.25 \times 3\,000 \times 25}{10^6} = 0.4 \text{ cycles per second/°C.}$$

For a tuned circuit at resonance, as is virtually true of the frequency-controlling element, the frequency : capacitance coefficient is given by

$$\Delta f/f = \Delta C_0/2C_0$$

which may be written as

$$\Delta f = 10^6/2C_0 \text{ cycles/megacycle/pF}$$

For the frequency : temperature coefficient, assuming the same temperature stability of capacitance, we obtain

$$\Delta f = \frac{10^6}{2C_0} \times C_0 \times \frac{25}{10^6} = 12.5 \text{ cycles per second/°C.}$$

Comparing the figure of 12.5 cycles per second/°C for the controlling capacitor with that of 0.4 cycle per second/°C for the input capacitance it will be seen that the effect upon frequency of the maintaining circuit due to change of temperature may be disregarded.

In addition to the effect of temperature variations, we have also to consider the stability of the valve interelectrode capacitances. The significant interelectrode capacitances are the grid-cathode capacitance and the cathode-earth capacitance, both of which are usually of the order of 10 pF. Experience shows that when a valve is left operating over a long period of time the variation of these capacitances does not usually exceed the order of 2%. Such a change in both capacitances simultaneously would, in the example quoted, cause the relatively insignificant frequency change of approximately 0.2 c/s.

So far we have considered the frequency

stability of the maintaining circuit in terms of variations in the value of the circuit capacitances which physically exist. We have referred briefly to the existence of what is in effect additional reactance, which does not exist physically, but results from a phase shift caused by the intermodulation of harmonic components generated by the non-linear characteristic of the valve. As a means of examining the effect of such phase shift, suppose that the phase-angle of the input impedance Z_g were caused to change slightly by adding to the valve current a small current with some arbitrary phase relationship. Since the real part of Z_g must always be equal in magnitude to the loss resistance R_0 of the controlling element, we may relate such a phase shift to a change in the effective value of C_g only, and from (12) we may deduce the resulting change in frequency.

In Appendix 3 this is shown to be given by:

$$\frac{df}{d\phi} = \frac{f}{2Q} + \frac{1}{\pi L_0 G} \quad .. \quad .. \quad .. \quad (16)$$

The first term on the right-hand side of this expression is the differential coefficient of frequency with respect to phase of a tuned circuit at resonance, while the second term is due to the effect of the maintaining circuit. Using the same values as in the previous examples, we find that only the second term is significant, this being approximately 100 times as great as the first. From this it will be seen that there is an advantage in using a high L/C ratio. It is not proposed to give any quantitative interpretation to the value of this coefficient, except to say that perhaps it is worth noting that the effect of an additional quadrature current flowing through Z_g with a relative amplitude of -60 db would be to cause a phase change of approximately 3.5 minutes of arc and a frequency change of 250 c/s. Such a current might well be caused by unintentional feedback from a subsequent amplifying stage, and if the relation between the signal fed back and the oscillation signal is not strictly linear, a relatively large frequency change may result, for example, from a small change in the supply voltage.

A direct comparison with the Colpitts circuit is simple and instructive. Since in this circuit a single inductance L_g is connected across C_g, we may from (5) and (9) immediately write the conditions

$$\omega = 1/\sqrt{(L_g C_g)} \quad .. \quad .. \quad .. \quad (17)$$

$$C_g \leqslant \frac{1}{2\omega}\sqrt{\left(\frac{G}{R_g}\right)} \text{ where } R_g = \frac{\omega L_g}{Q} \quad .. \quad (18)$$

Substituting (17) in (18) we obtain

$$C_g = L_g G/4R_g = QG/4\omega \quad .. \quad .. \quad (19)$$

For a resonant circuit we know that $\Delta f/f = \Delta C/2C$ whence

$$\frac{\Delta f}{f} = \frac{4\pi f \Delta C_g}{QG} \quad .. \quad .. \quad .. \quad (20)$$

which is precisely the same as for the modified circuit as expressed in (13). This is to be expected since we have already shown that for the modified circuit this coefficient is independent of the L_0/C_0 ratio, and the Colpitts circuit is in effect the modified circuit with the L_0/C_0 ratio put to zero.

However, when we examine the value that C_g would have in a practical case we find from (19) using the previous values, $\omega = 2\pi \times 10^6$, $Q = 200$, $G = 6$ mA/V

$$C_g = \frac{200 \times 6 \times 10^{12}}{4 \times 10^3 \times 2\pi \times 10^6} = 48000 \text{ pF}$$

while L_g would be 0.53 μH.

The impracticability of using such values is obvious, but the comparison nevertheless demonstrates that the stability of the Colpitts circuit is theoretically as great as that of the modified circuit insofar as component changes are concerned. We have, however, shown that the effect upon frequency of a phase change caused by non-linearity is directly proportional to the L/C ratio, and in this respect the modified circuit will be of the order of 150 times as stable.

3. The Amplitude Limiter

If an oscillator of the type described were to be constructed with circuit values chosen accurately for operation at a frequency of the order of 1.0 Mc/s, the measured frequency would be found to differ from the exact design frequency by some 100 to 500 cycles per second according to the amplitude of oscillation. If the amplitude of oscillation were to be reduced by any means, for example by reducing the valve current, the discrepancy would decrease and would in fact tend to zero as the amplitude of oscillation approached zero. A complete demonstration of this fact is difficult to perform, since at small amplitudes of oscillation a valve characteristic is too linear to provide a condition of stable equilibrium, but it is only too easy to demonstrate with any simple oscillator that the frequency is to some extent a function of the supply voltages!

As mentioned earlier, the phenomenon may be explained in terms of the phase shift which results from intermodulation products, but there is also a simple mechanical analogue, which, although not in all respects perfect, may nevertheless be found instructive.

Consider a freely suspended pendulum, the free period of oscillation of which has been accurately deduced from its known constants. Having displaced the pendulum from rest and allowed it to swing, we find that in order to maintain it in oscillation at a constant amplitude we must, after each return stroke, tap the bob so as to replace the

energy that has been lost over the cycle due to friction. The action of tapping the bob must necessarily alter its velocity, since if this were not so, no force would be applied, and if while so maintained in oscillation, the frequency of the oscillation were measured, it would be found to be somewhat greater than the calculated frequency due to the increased velocity of movement on the forward stroke. The error could of course be corrected by adjusting the effective length of the pendulum, but the frequency would then only be exact for one amplitude of oscillation corresponding to one strength of 'tap.' The smaller the amplitude of swing, the more closely will the displacement of the bob follow simple harmonic motion, and the more accurately will the maintained frequency equal the 'free' frequency of the pendulum.

If we wish to construct an oscillator to have the maximum possible frequency stability, we must, therefore, include some form of limiter which will provide a condition of stable equilibrium when the actual amplitude of oscillation is small—sufficiently small to be confined to a substantially linear portion of the valve characteristic. A simple and effective method of limiting the oscillation is shown by the block schematic in Fig. 5. The signal output of the oscillator valve is amplified, rectified, and the d.c. signal so obtained is applied as a bias voltage to the oscillator grid.

By this means it is a simple matter to arrange that the oscillation will limit at any desired level.

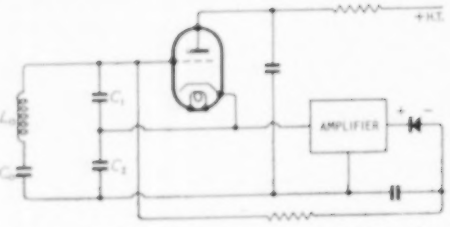

Fig. 5. *Simple oscillation amplitude-limiting circuit.*

For example, suppose that a slope of 6 mA/V is required of the oscillator valve to maintain the oscillation, and that from the valve characteristic we find that for a given anode voltage a grid bias of −3.0 volts will provide this slope. If we wish the oscillation to limit at a peak amplitude of, say, 0.1 V, measured between grid and cathode, an amplification of 30 will provide the required bias voltage. The amplified signal is of course available as an output signal, and in the example quoted above a 3.0-V peak signal would be obtained at the input to the rectifier.

As with most devices which provide automatic control by means of feedback, the circuit constants must be chosen to avoid instability. Instability in the form of a low-frequency relaxation oscillation will result if the time constant of the r.f. filter following the rectifier is too long, but it is not usually difficult to choose values that will provide adequate attenuation to the r.f. signal without introducing appreciable delay to the rectified control signal.

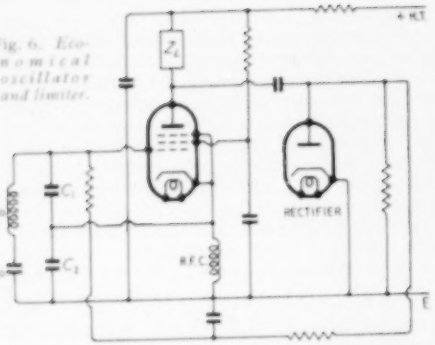

Fig. 6. *Economical oscillator and limiter.*

In one experiment, the author constructed a limiting circuit with a gain sufficient to cause the amplitude of oscillation to limit at a peak voltage of the order of 0.01 V. At such a small amplitude the harmonic distortion was, of course, negligible, and in order to prove that the frequency of oscillation was then precisely equal to the loop resonance of the controlling circuit and substantially unaffected by the valve parameters the oscillation was gradually reduced to zero amplitude by

(a) reducing the anode voltage;
(b) reducing the heater voltage;
(c) adding a negative bias voltage to the control grid.

In each case the frequency did not change by more than one part in 10^6 before the oscillation ceased completely. It will be appreciated that a *low* amplitude of oscillation does not necessarily mean a 'weak' oscillation in the accepted sense of the term. The 'strength' of the oscillation is indicated by the built-up time, and depends upon the extent to which the energy available at the instant of switching on exceeds the energy required to maintain the oscillation.

A very simple and economical arrangement which will provide a useful control of the amplitude is shown in Fig. 6. By making V_1 a pentode, we may derive an amplified signal from the anode without seriously disturbing the operation of the circuit, provided the anode-load impedance, shown as Z_L, is not excessive. In practice, a gain of about 10 may be obtained without difficulty, and this is sufficient to operate the limiter to give

a worth-while improvement in performance.

In a superior form of limiter, a separate valve is used to provide the necessary gain, in which case the oscillator valve may be connected as a triode with the anode decoupled directly to earth, and the signal output may be conveniently taken from the cathode, at which point, as has been shown, the impedance is very low.

4. Practical Performance

A practical form of the oscillator complete with limiter is shown in Fig. 7. For high-stability the series LC controlling circuit together with the two capacitors C_1 and C_2 are mounted in a temperature-controlled compartment or oven. Since the impedance measured at the grid and cathode is very low, stray capacitance introduced by the connections, which may be fairly long, is relatively unimportant.

If it is required to vary the frequency of the oscillator, the capacitors C_1 and C_2 must of course be chosen for the highest frequency, in which case the amplitude of the oscillation will increase as the frequency is reduced due to increased coupling.

For high stability over a wide frequency band it is, therefore, desirable to divide the total range into a number of smaller ranges; for example an octave may be conveniently divided into three ranges each of ratio approximately 1.26 : 1. The components L_0, C_0, C_1 and C_2 are then chosen for each range separately and a switching arrangement provided so that any one of the three complete assemblies may be connected to the oscillator valve. It will be noted that the switching takes place at the grid and cathode connections to the valve where the impedance looking towards the tuning elements is very low, while looking into the valve it is relatively high, so that neither stray shunt capacitance or stray series impedance will have a serious effect upon the frequency of oscillation.

A circuit of this form has been used for several years in B.B.C. short-wave transmitter-drive equipment with considerable success. In the actual circuit in use two high-slope pentodes are connected in parallel in order to obtain twice the mutual conductance of a single valve, thereby permitting the impedance measured across the series-tuned circuit to be approximately halved.

Below are quoted some measured performance data for this type of oscillator:

(1) H.T./frequency coefficient $\pm 2.5/10^6$ for $\mp 20\%$ h.t. voltage

(2) L.T./frequency coefficient $\pm 2.5/10^6$ for $\mp 5\%$ l.t. voltage

(3) Frequency Stability

(a) Short term (about 1 hour) $\pm 1/10^6$

(b) Med. term (about 1 day) $\pm 10/10^6$

(c) Long term (about 1 month) within re-setting accuracy of $\pm 30/10^6$

Needless to say, it was necessary to pay great attention to mechanical design and temperature stabilization in order to obtain this stability, but the results serve to illustrate the capabilities of the circuit.

As mentioned earlier, the circuit has been used with equal success as a crystal-controlled oscillator. Since the equivalent electrical circuit of a crystal approximates to a series LC circuit, it is only necessary to replace the series LC elements with a crystal for this purpose. A high-grade crystal cut for medium frequencies and carefully mounted usually has an equivalent series resistance of the order of 100 Ω and the appropriate value of the capacitors C_1 and C_2 is therefore considerably less than for an LC controlled circuit, for which the series resistance

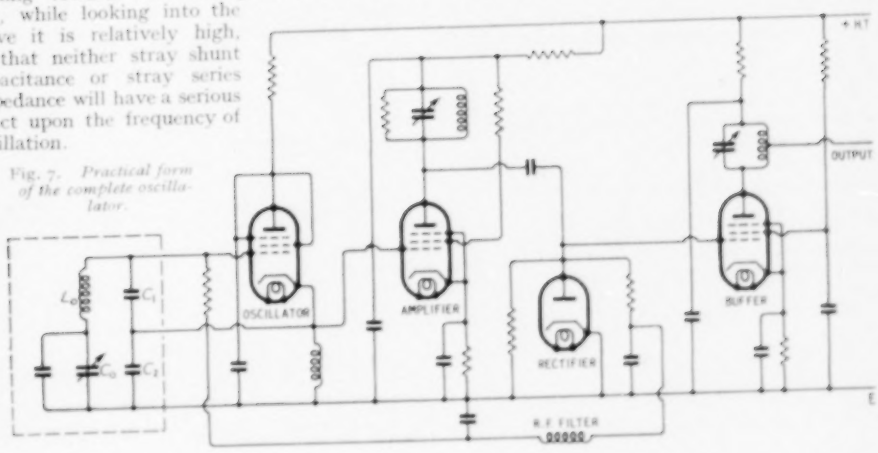

Fig. 7. Practical form of the complete oscillator.

is usually much less than 10 Ω. However, due to the high Q of a crystal as compared with an LC circuit, the effect upon frequency of a small change in the values of these capacitances is exceedingly small—in practice of the order of $2/10^8$ c/s per pF change. Again, in such circumstances it is doubtful if the maintaining circuit will contribute to any measurable extent to the overall instability of the oscillation.

5. Acknowledgment

The author wishes to thank the Chief Engineer of the British Broadcasting Corporation for granting permission to publish this article.

APPENDIX 1

Input Impedance

The input impedance of the circuit will be analysed with reference to Fig. 4(b).

Assuming the valve to have an infinite impedance (e.g., a perfect pentode) we may write —

Current flowing through $Z_2 = I_g + e_g g_m$ . . (1a)

Voltage across $Z_2 = Z_2 (I_g + e_g g_m) = e_g$ . . (2a)

and as $E_g = e_g + e_g$

$$E_g = Z_2 (I_g + e_g g_m) + e_g$$
$$= I_g Z_2 + e_g (1 + g_m Z_2) \quad . . \quad . . \quad (3a)$$

but $e_g = I_g Z_1$

$$\therefore E_g = I_g Z_2 + I_g Z_1 (1 + Z_2 g_m) \quad . . \quad . . \quad (4a)$$

whence $E_g = Z_g = Z_1 + Z_2 + g_m Z_1 Z_2$ . . (5a)

We have assumed a perfect pentode but in fact it is only necessary that the valve impedance should be much greater than Z_2. Since, as is shown in the text, the impedance of Z_2 is, in practice, of the order of 100 Ω, the analysis is valid for any normal triode.

APPENDIX 2

Evaluation of the Coefficient $\frac{df}{dC_g}$

Solving (5) for frequency, we obtain

$$f = f_0 \sqrt{(1 + C_0/C_g)} \quad . . \quad . . \quad . . \quad . . \quad (6a)$$

where f_0 is the resonant frequency of $L_0 C_0$.

Differentiating (6a) with respect to C_g we obtain

$$\frac{df}{dC_g} = -f_0 \cdot \frac{C_0}{C_g^2} \cdot \frac{1}{2\sqrt{(1 + C_0/C_g)}} \quad . . \quad . . \quad (7a)$$

and since C_0 is small compared with C_g, and also f_0 is very nearly equal to f, we have approximately

$$\frac{df}{dC_g} = -\frac{fC_0}{2C_g^2} \quad . . \quad . . \quad . . \quad . . \quad (8a)$$

Also since $\omega L_0 \approx 1/\omega C_0$ we may write

$$Q = \frac{\omega L_0}{R_0} = \frac{1}{\omega C_0 R_0} \text{ or } R_0 = \frac{1}{\omega C_0 Q} \quad . . \quad . . \quad (9a)$$

By substituting this value for R_0 in (9) and taking the maximum permissible value of C_g we obtain

$$C_g = \frac{1}{2\omega} \sqrt{GQ\omega C_0} \quad . . \quad . . \quad . . \quad . . \quad (10a)$$

and therefore $\frac{C_0}{C_g^2} = \frac{4\omega}{QG}$ . . . . . (11a)

From (8a) and (11a) we therefore have

$$\frac{df}{dC_g} = -\frac{4\pi f^2}{QG} \quad . . \quad . . \quad . . \quad . . \quad (12a)$$

APPENDIX 3

Evaluation of the Coefficient $\frac{df}{d\phi}$

The phase angle of the input impedance, Z_g, is given by $\phi = \tan^{-1} \frac{1}{\omega C_g R_g}$ . . . . . (13a)

as shown in Fig. (8)

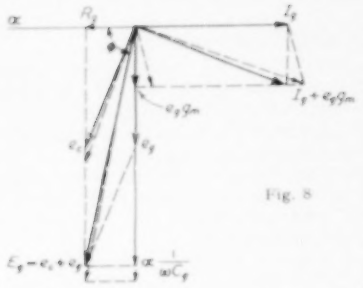

Fig. 8

Since R_g is at all times numerically equal to the loss resistance R_0 (but of opposite sign) any change in the phase angle ϕ must be accounted for by a change in the reactive component only, and it is therefore equivalent to a change in the value of C_g. This is illustrated by the dotted vectors in Fig. (8) which indicate the effect of a phase displacement originating in the valve current vector.

From (13a) we obtain by differentiation

$$\frac{dC_g}{d\phi} = -\left[\frac{1}{\omega R_g} + \omega R_g C_g^2\right] \quad . . \quad . . \quad . . \quad (14a)$$

or $\frac{dC_g}{d\phi} = \frac{1}{\omega R_0} + \omega R_0 C_g^2$ . . . . . (15a)

since $R_g = -R_0$

From (9) we obtain

$$C_g^2 = \frac{1}{4\omega^2} \frac{G}{R_0} \quad . . \quad . . \quad . . \quad . . \quad (16a)$$

which by substitution in (15a) gives

$$\frac{dC_g}{d\phi} = \frac{1}{\omega R_0} + \frac{G}{4\omega} \quad . . \quad . . \quad . . \quad . . \quad (17a)$$

Also from (12) we have

$$\frac{df}{dC_g} = \frac{4\pi f^2}{QG} \quad . . \quad . . \quad . . \quad . . \quad (18a)$$

whence

$$\frac{df}{d\phi} = \frac{dC_g}{d\phi} \cdot \frac{df}{dC_g}$$
$$= \frac{4\pi f^2}{QG} \left[\frac{1}{\omega R_0} + \frac{G}{4\omega}\right] \quad . . \quad . . \quad . . \quad (19a)$$
$$= \frac{2f}{QGR_0} + \frac{f}{2Q} \quad . . \quad . . \quad . . \quad . . \quad (20a)$$

and since $Q = \omega L_0/R_0$

$$\frac{df}{d\phi} = \frac{1}{\pi L_0 G} + \frac{f}{2Q} \quad . . \quad . . \quad . . \quad . . \quad (21a)$$

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1 "An Inductance-Capacity Oscillator of Unusual Frequency Stability," J. K. Clapp, *Proc. Inst. Radio Engrs*, March 1948, Vol. 36, No. 3.

2 For example, see *Q.S.T.*, May 1948, p. 42, August 1948, p. 28, January 1949, p. 43.

3 "Constant-Frequency Oscillators," Llewellyn, *Proc. Inst. Radio Engrs*, 1941, Vol. 19, pp. 2063–2094.

IMPEDANCE MATCHING NETWORKS

By R. O. Rowlands, B.Sc.

(B.B.C. Engineering Training Department)

SUMMARY.—A method is described of designing a four-terminal network whose image impedance at one pair of terminals is a constant resistance and at the other pair of terminals a complex quantity varying with frequency.

Introduction

IT is sometimes necessary to connect equipment, such as an unloaded cable whose characteristic impedance W'' is a complex quantity varying with frequency, to some other equipment whose impedance is nearly a constant resistance R over the working frequency band. To overcome the bad effects of reflection due to connecting two such dissimilar impedances together an impedance-matching network is inserted between the two pieces of equipment. This network is designed to have image impedances W and R; W being nearly equal to W'' over the frequency band.

Let the network of Fig. 1 (a) fulfil these conditions. If two such networks are connected back to back at the terminals whose impedance is W, then the resulting network is a symmetrical constant-resistance network and so may be transformed into a lattice whose series and lattice arms Z_1 and Z_2 are inverse with respect to R; i.e., $Z_1 Z_2 = R^2$

The most general method of design is therefore to start with a constant-resistance lattice network, expand it into a ladder network and bisect it.

If the lattice is capable of being expanded into a ladder network, then the arms Z_1 and Z_2 will contain a common series or shunt impedance. Let the former be the case, then

$$Z_1 = Z'_1 + Z_a$$

and

$$Z_2 = Z'_2 + Z_a$$

The common impedance Z_a can be taken outside the lattice in series with both the input and the output as shown in Fig. 1 (b).

The remaining arms Z'_1 and Z'_2 will now have a common shunt impedance Z_b and this can be taken outside the network in shunt with the input and output as shown in Fig 1 (c).

This process may be repeated taking out successive common series and shunt impedances until one pair of arms becomes either infinite or zero. The resulting ladder network is symmetrical and so may be bisected down the middle.

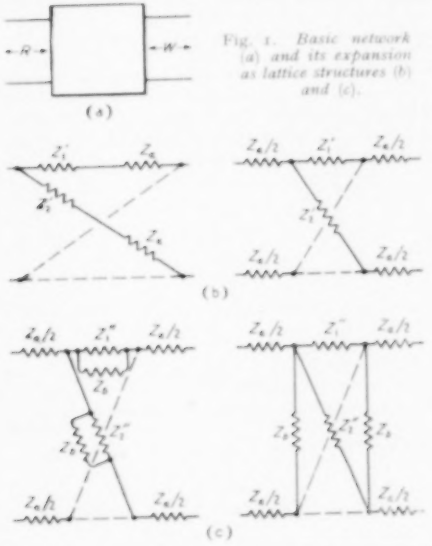

Fig. 1. *Basic network (a) and its expansion as lattice structures (b) and (c).*

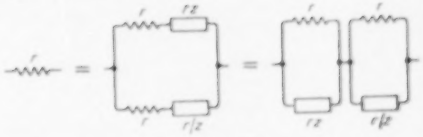

Fig. 2. *Equivalent networks, types 1 and 2 respectively.*

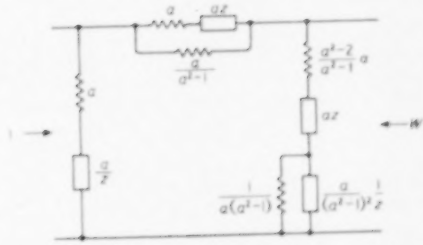

Fig. 3. *Network type 1.2.*

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An important relation which will be used in all the succeeding expansions is that a resistor r may always be replaced by either of the networks of Fig. 2 where rz is any two-terminal network containing resistors, inductors, and/or capacitors, and r/z is its inverse with respect to r.

Impedance-Matching Networks

Type 1.1

The simplest type of constant-resistance lattice network which can be expanded as a ladder structure is one containing a single resistor in each arm. The straightforward expansion of this

TABLE 1

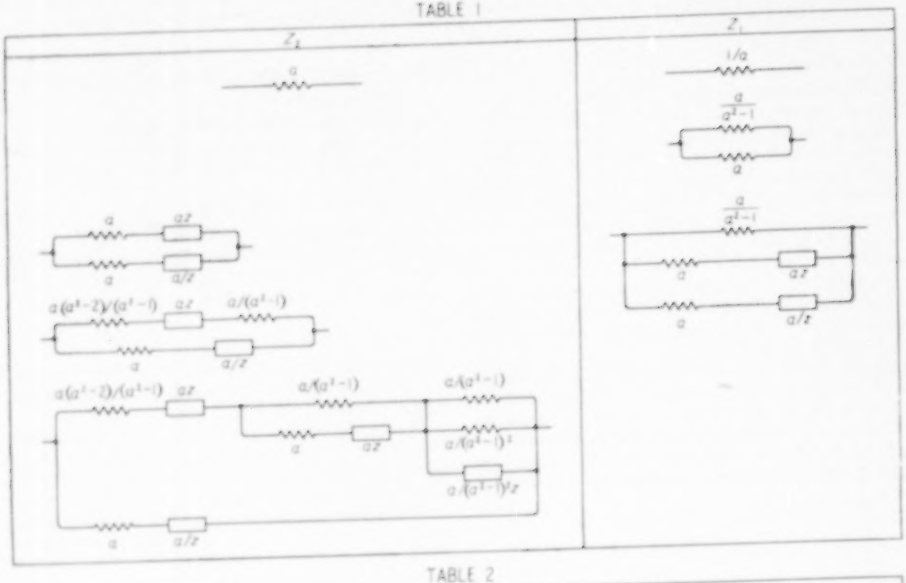
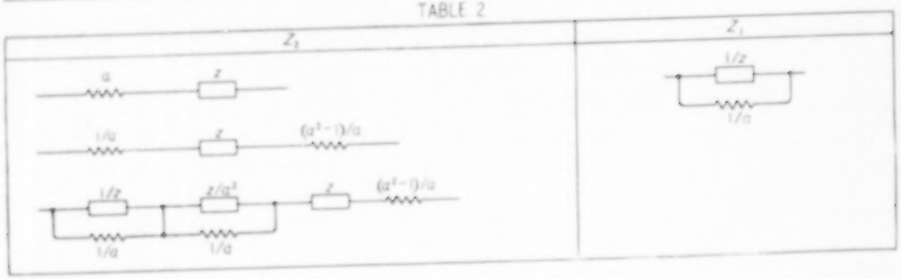


TABLE 2



Explanation of Table 1.

Step 1. Z_1 *is split into two resistors in parallel one of which has the value* 'a'.

Step 2. *The resistors* 'a' *in both* Z_1 *and* Z_2 *are replaced by an equivalent network, type 1.*

Step 3. *The resistor* 'a' *of* Z_2, *which is in series with* 'az' *is split into two resistors in series, one of which is equivalent to the shunt resistor of* Z_1.

Step 4. *This is replaced by an equivalent network type 2 part of which is equal to the two upper shunt arms of* Z_1.

Explanation of Table 2.

Step 1. *The resistor* 'a' *of* Z_2 *is split into two resistors in series one of which has the same value as the shunt resistor of* Z_1.

Step 2. *This resistor in* Z_2 *is replaced by an equivalent network type 2 part of which is the same as the whole of* Z_1.

network leads to the trivial case of an attenuator pad which matches the constant resistance R to another constant resistance nR.

Type 1.2

In this and in subsequent networks the value of R will be taken as unity for simplicity. If Z_1 and Z_2 are again purely resistive then each of them may be built out as shown in Table 1 by successive applications of the equivalent-network relationships of Fig. 2. If the value of Z_2 is a then the value of Z_1 is $1/a$.

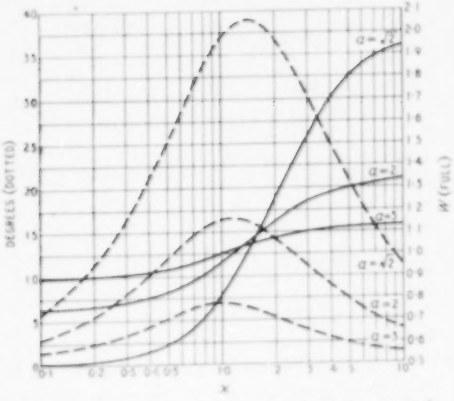


Fig. 4. *Modulus and argument of W for network type* 1.2 *when* $z = jx$.

The impedance-matching network obtained by expanding and bisecting the lattice is shown in Fig. 3.

For this network to be physically realizable the value of a must be greater than or equal to $\sqrt{2}$. Since it is derived from pure resistors it gives a constant loss at all frequencies.

The impedance W is given by :—

$$W = \frac{a^2(a^2-1)(1+z)^2}{\{a^2+(a^2-1)z\}^2} = \left|\frac{1+z}{\frac{a}{\sqrt{(a^2-1)}}+\frac{\sqrt{(a^2-1)}}{a}\cdot z}\right|$$

If $\dfrac{a}{\sqrt{(a^2-1)}} = b$ and $z = jx$

Then

$$W = \frac{(1+jx)^2}{(b+jx/b)^2} = \frac{(b-jx/b+jbx+x^2/b)^2}{(b^2+x^2/b^2)^2} = \frac{\{(b+x^2/b)+jx(b-1/b)\}^2}{(b^2+x^2/b^2)^2}$$

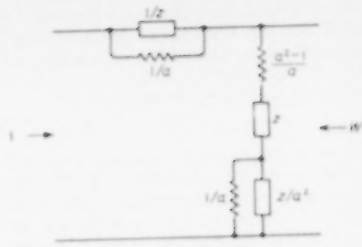


Fig. 5. *Network type* 2.1.

Therefore Mod.

$$W = \frac{(b+x^2/b)^2+x^2(b-1/b)^2}{(b^2+x^2/b^2)^2} = \frac{b^2+2x^2+x^4/b^2+b^2x^2-2x^2+x^2/b^2}{(b^2+x^2/b^2)^2} = \frac{(b^2+x^2/b^2)(1+x^2)}{(b^2+x^2/b^2)^2} = \frac{1+x^2}{(b^2+x^2/b^2)}$$

and Arg.

$$W = 2\tan^{-1}\frac{x(b-1/b)}{b+x^2/b} = 2\tan^{-1}\frac{x(b^2-1)}{b^2+x^2}$$

The modulus and argument of W for the type 1.2 network are shown in Fig. 4 for several values of a.

Types 1.3, *etc.*

More complicated networks may be built up by developing the arms Z_1 and Z_2 in different ways ; e.g., the resistance a in series with az in the arm Z_2 may be divided into two parts,

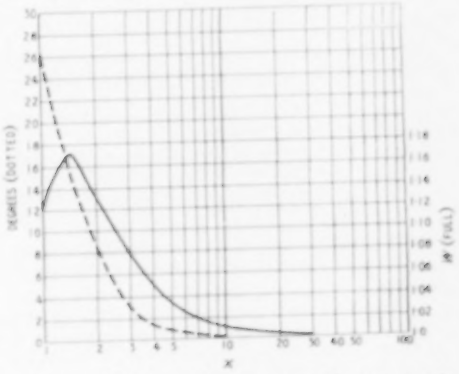


Fig. 6. *Modulus and argument of W for network type* 2.1 *when* $z = jx$ *and* $a = 1$.

one of which is less than $a/(a^2-1)$. This is replaced by its equivalent network as before and then the arm Z_1 is developed a stage further in such a way that the whole network may be built out as a ladder structure.

Type 2.1

A constant-resistance lattice structure which is a step more complicated than the one with a single resistor in each arm is one containing a resistor in series with a network z in one arm and a resistor in parallel with a network $1/z$ in the other arm. Each arm can be built out as shown in Table 2 and from this is obtained the impedance matching network of Fig. 5.

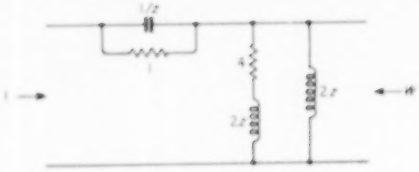


Fig. 7. *Special case of type 2.1.*

For this network

$$W = \frac{(1+a+z)(a^2-1+2az+az^2)}{(a+z)(a^2+a+z+2a^2z+az^2)}$$

The modulus and argument of W are shown in Fig. 6 for $z = jx$ and $a = 1$.

In the special case of the above where $a = 1$ and z is an inductor then the network may be built as shown in Fig. 7.

An advantage of this network is that the inductor appearing in shunt across the right-hand terminals may be wound as a transformer, thus altering the original impedance transformation by a constant multiple.

Type 2.2

The same basic lattice arms may be built up into more complicated structures as shown in Table 3. The resulting impedance-matching network is shown in Fig. 8. The expression for the impedance W is rather complicated and is best worked out for particular values of a, b, and z.

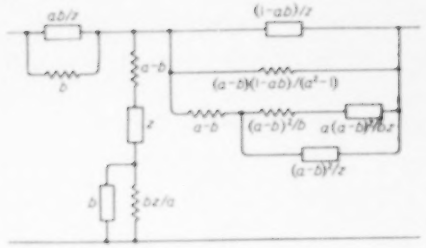


Fig. 8. *Network type 2.2.*

For the network to be physically realizable we must have :—

$a > b$ 1

$1 > ab$ 2

$a^2 > 1$ 3

i.e.,

$a > 1 > 1/a > b$

Type 3.1

The basic network of this type contains two resistors in each arm. It may be developed as shown in Table 4, resulting in the impedance-matching network of Fig. 9.

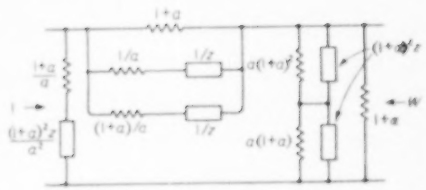


Fig. 9.—*Network type 3.1.*

Explanation of Table 3 (opposite).

Step 1. *The resistor 'a' of Z_2 is split into two resistors in series one of which has the value 'b', and Z_1 is split into two parts one of which contains a resistor 'b'.*

Step 2. *The resistor 'b' of Z_2 is replaced by an equivalent network type 2 part of which is the same as that part of Z_1 which contains the resistor 'b'. The other resistor in Z_1 is replaced by two resistors in parallel one of which has the value '$a - b$'.*

Step 3. *The resistor '$a - b$' is replaced by an equivalent network type 1, part of which is the same as that part of Z_2 which was not previously common to Z_1.*

Explanation of Table 4 (opposite).

Step 1. *The whole of Z_2 is replaced by an equivalent network with a shunt resistor. The resistor 1 in Z_1 is replaced by two resistors in parallel one of which is equivalent to the shunt resistor of Z_2.*

Step 2. *The other parallel resistor of Z_1 is replaced by an equivalent network type 1, one arm of which is the same as that of Z_2.*

Step 3. *The resistor $1+a$ of Z_2 is replaced by an equivalent network type 2 part of which is the same as the whole of Z_1 except for the arm mentioned in step 2.*

TABLE 3

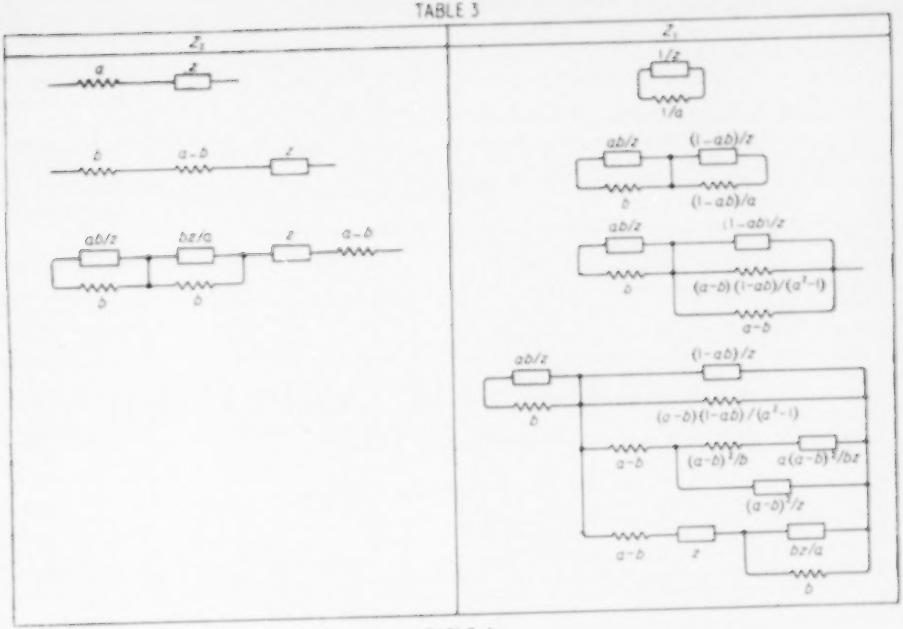
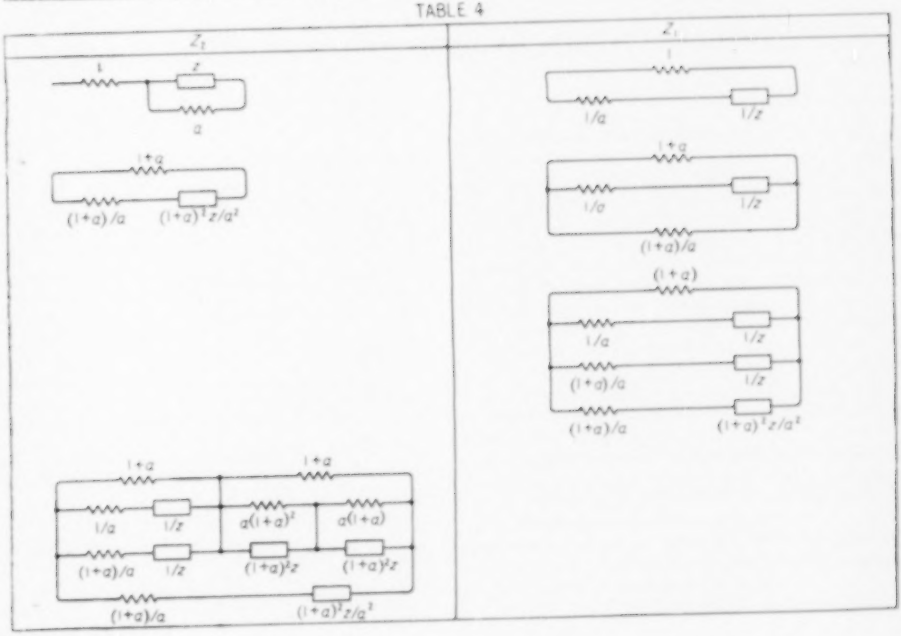


TABLE 4



The impedance W of the above network is given by

$$W = \frac{s\,(t - u)}{s - t - u}$$

Where

$$s = \frac{a\,(1-a)\,z\,\{2a - (2-a)\,z\}}{(1-4a-2a^2)z^2 - a(2-5a-a^2)z - a^2(1-a)}$$

$$t = \frac{(1-a)\,(a-z)\,\{a - (1-a)\,z\}}{(1-a)^2\,z^2 - (2-3a-2a^2)\,az - a^2}$$

$$u = \frac{(1-a)\,\{a - (1-a)\,z\}}{(1-a)^2 z - a\,(1-2a)}$$

Further Types

There is no limit to the number of types of networks that may be constructed by the method outlined, but they will be successively more complex and will eventually cease to be of any practical value.

To each of the networks already described there also exists a dual which may be obtained from the original by the standard method of constructing dual networks*.

Acknowledgment

The author is indebted to Dr. Sturley for reading the manuscript of the paper.

* A method of doing this is described in Guillemin's "Communication Networks," Vol. II, page 246.

THE MAGNETIC AMPLIFIER

Transductor Theory

By R. Feinberg, Dr.Ing.

(Electrical Engineering Dept., University of Manchester)

SUMMARY.—The transductor is the vital element of a magnetic amplifier circuit. The conditions for obtaining optimum linear control of a transductor with free magnetization with no load and sinusoidal voltage are: a transductor core having a magnetization curve with a sharp bend at the knee and a high initial permeability, and a transductor voltage of such magnitude that the peak value of the alternating magnetic-flux density in the transductor core is equal to the value of flux density at the knee of the magnetization curve. The control sensitivity on the linear part of the transductor characteristic is substantially equal to the turns ratio between control winding and load winding of the transductor, apart from a factor of 2, accounting for parallel connection of the transductor load coils in the case of the parallel-type transductor.

Introduction

MAGNETIC amplifiers have in recent years increasingly been coming into practical use for special applications as reliable substitutes for vacuum-valve amplifiers.[1] The principle upon which a magnetic amplifier is built has been known for about fifty years and has in various forms been utilized in heavy-current control and measuring techniques.[2] It was, for an experimental period, also applied in the technique of wireless communication,[3] and there is at present an interest in its use, in modified form, in high-frequency technique.[4]

The transductor is the vital element of a magnetic amplifier circuit. It is the purpose of this paper to present the theory of a transductor characteristic for practical evaluation.

Basic Circuit of Magnetic Amplifier

Fig. 1 shows the basic arrangement of a magnetic amplifier circuit. A source 1 of alternating current of fixed voltage V_1 supplies current, I_2, to a load 2 which is connected in series with a transductor 3 containing a 'load winding' 3a and a 'control winding' 3b, the control winding carrying the control current, I_c supplied from a direct-current source 4 of variable voltage V_c. 'Transductor' is the short technical term for a ferromagnetic-core inductor with variable direct-current bias-magnetization.

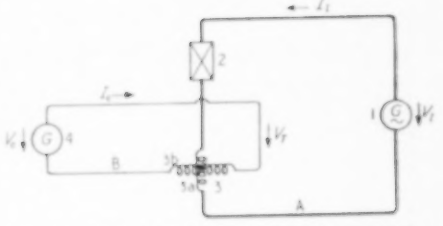


Fig. 1. Basic circuit arrangement of a magnetic amplifier: A = Load circuit, B = Control circuit, 1 = Alternating-current source of fixed voltage, 2 = Load, 3 = Transductor, 3a = Load winding, 3b = control winding, 4 = Source of control current

The graphical symbol used in Fig. 1 to denote a transductor shows two windings crossing one

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another at right-angles. The symbol means that the two windings are placed on a common core but have no transformer interaction.

The magnetic amplifier, Fig. 1, is operated by varying the control current I_c and obtaining a corresponding change of the load current I_L. The transductor acts in the load circuit as a variable-impedance element controlled by I_c, and is so designed that a small change of I_c causes a magnified change of I_L.

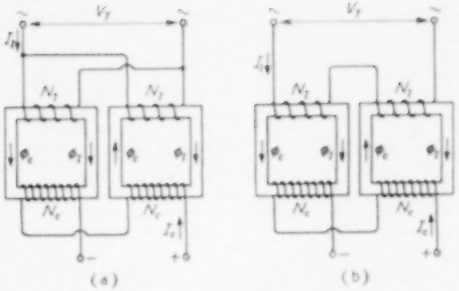


Fig. 2.—*Transductor with electric elimination of transformer interaction between load winding and control winding, (a) parallel-type and (b) series-type.*

Basic Transductor Structure

A transductor is a modification of an iron-core transformer. Transformer interaction between load winding and control winding is avoided by using either an electric or a magnetic method.

Fig. 2 illustrates the electric method of avoiding transformer interaction. The transductor consists of two 2-limb cores (or ring-shaped cores) with the control winding and the load winding split into two coils each. The two control coils are electrically connected in series-opposition with the result that the fundamental components of the e.m.fs induced in them cancel one another in the control circuit. The two coils of the load winding are connected either in parallel, see Fig. 2(a), or in series, see Fig. 2(b).

The magnetic method of avoiding transformer interaction is shown in Fig. 3. A 3-limb core carries the control winding as a single coil on the middle limb and the load winding as two coils on the outer limbs the two load coils being connected either in parallel, as in Fig. 3(a), or in series, as in Fig. 3(b). The fundamental components of the alternating magnetic fluxes excited by the load coils are in series-opposition in the middle limb of the core and cancel one another, the control circuit is thus free from a corresponding e.m.f.

In a practical transductor either the electric method of Fig. 2 or the magnetic method of Fig. 3, or a modification of either of the two, is used for avoiding transformer interaction.

Mechanism of Transductor Operation

The control mechanism of a transductor is based on the magnetic saturability of the core material. Figs. 4 and 5 demonstrate the effect of direct-current bias-magnetization on the shape of the magnetization loop of a transductor core made of Permalloy C. The oscillograms of Fig. 4 were taken with a large-amplitude sinusoidal voltage at the transductor load winding and the oscillograms of Fig. 5 with a small-amplitude voltage. Fig. 4 shows that increasing bias-magnetization makes the magnetization loop increasingly unsymmetrical, and Fig. 5 illustrates how the magnetization loop is "lifted" over the knee of the magnetization characteristic into the upper region of magnetic saturation.

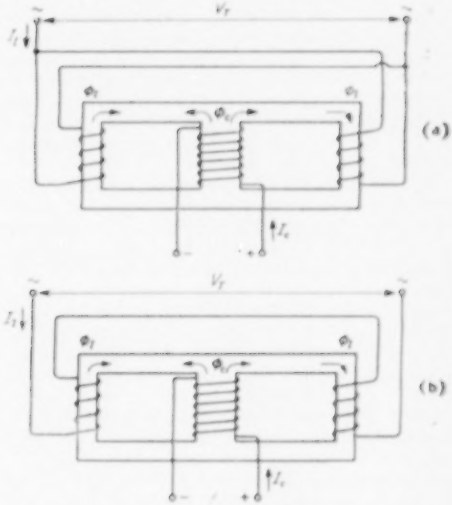


Fig. 3. *Transductor with magnetic elimination of transformer interaction between load winding and control winding, (a) parallel-type, (b) series-type.*

The change of the magnetization loops in Figs. 4 and 5 means in terms of the circuit arrangement of Fig. 1 that an increase of control current I_c causes an increase of load current I_L. The quantitative relation between the changes of I_c and I_L is determined by the electric characteristic of the transductor and the impedance of the load, and by the mode of transductor magnetization.

There are two principal modes of transductor magnetization, respectively denoted free magnetization and constrained magnetization.[5] In a transductor with free magnetization part of the a.c. magnetization current flows outside the load circuit (A in Fig. 1) so that the load current, I_L, is smaller than the total a.c. magnetization current of the transductor. In the case of constrained magnetization I_L is identical with the total a.c. magnetization current.

Fig. 4 (left). *Large-amplitude magnetization loops of a Permalloy C core and varying direct-current bias-magnetization. (a) No bias-magnetization. (b) Slight bias magnetization. (c) Bias-magnetization increased.*

We have, for example, free magnetization in the parallel-transductor (see Figs. 2(a) and 3(a)) with sinusoidal voltage, V_T. The even current harmonics of a.c. magnetization form a circulating current in the two coils of the load winding and are thus absent from the load circuit; i.e., I_L is smaller than the total a.c. magnetization current.

The series-transductor (see Figs. 2(b) and 3(b)) with sinusoidal voltage V_T operates either in free or in constrained condition, or somewhere in between. The criterion for the condition of operation is given by the impedance of the load circuit (B in Fig. 1) in relation to the impedance of the load circuit (A in Fig. 1). Suppose that the control circuit impedance is very small: each transductor core has sinusoidal a.c. magnetization with the even current harmonics required flowing freely in the control circuit. Part of the a.c. magnetization current thus flows outside the load circuit; i.e., the load current, I_L, is smaller than the total a.c. magnetization current. We have free condition. In the other extreme, assume that the impedance of the control circuit is very high so that no alternating current component can flow in the control circuit: the a.c. magnetization of each transductor core is forced to be non-sinusoidal. The load current, I_L, provides the total a.c. magnetization; i.e., the transductor operates in condition of constrained magnetization.

Transductor Characteristic

The electrical characteristic of a free transductor depends on the waveforms of transductor voltage and load current. The voltage is sinusoidal, like the supply voltage, when the transductor has no load (i.e., when in Fig. 1 the load 2 is short-circuited); the load current contains odd harmonics and is, therefore, non-sinusoidal. On the other hand, we have sinusoidal load current and non-sinusoidal transductor voltage when the transductor is in series with a resistive or inductive load of high impedance. The characteristics differ from one another in the two extreme conditions of waveforms. In normal load condition the transductor characteristic is intermediate between the two extremes.

Fig. 5. *Small-amplitude magnetization loops of the transductor used for Fig. 4. (a) No bias-magnetization. (b) Slight bias-magnetization. (c-f) Bias-magnetization increased in stages.*

It is proposed to calculate the characteristic for sinusoidal transductor voltage. The assumptions for calculation are a transductor of Fig. 2 with free magnetization, sinusoidal transductor voltage V_T (see Figs. 1 and 2), a simplified magnetization curve consisting of straight lines (see Fig. 6), perfect magnetization of the core (i.e., no leakage flux) and disregard of iron and copper losses in the transductor.

Fig. 7 illustrates the basis of calculation. The transductor voltage

$$v_T = -V_T\sqrt{2}\sin\omega t \qquad (1)$$

and the bias-magnetization current I_c produce, in each transductor core, Fig. 2, a magnetic flux density of instantaneous value

$$b = B_p + B_r\sqrt{2}\cos\omega t \qquad (2)$$

where B_p is the mean value of magnetic-flux

density in a core and B_r is related to V_T by the equation

$$B_r = k_r \frac{V_T}{\omega A N_l} \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

with $k_r = 1$ for Fig. 2(a) and $k_r = 0.5$ for Fig. 2(b); A denotes the cross-section of a core, $\omega = 2\pi f$ the angular frequency of the supply voltage, and N_l designates the number of turns of a load coil.

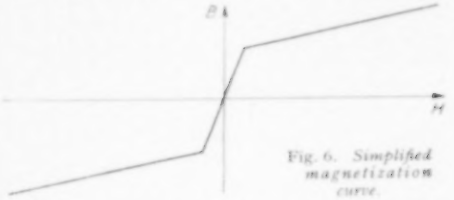


Fig. 6. *Simplified magnetization curve.*

B_p is related to I_c and B_r. Let H_0 be the mean value of magnetizing force in a transductor core, l the mean length of magnetic path in a core assumed to be of uniform cross-section A, and N_c the number of turns of a control coil. Then

$$H_0 = I_c N_c / l \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

The magnetizing force, H_p, corresponding to B_p (see Fig. 7), is less than H_0 because the convex shape of the magnetization curve produces a magnetic rectification effect of a negative sign. To evaluate H_p and hence B_p in terms of H_0, see equation (4), and of B_r is of no direct interest for the calculation of the transductor characteristic.

To facilitate calculation of the transductor characteristic it is proposed to consider B_p as

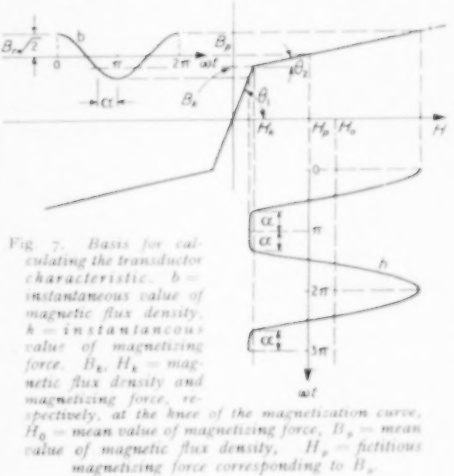


Fig. 7. *Basis for calculating the transductor characteristic. b = instantaneous value of magnetic flux density, h = instantaneous value of magnetizing force. B_k, H_k = magnetic flux density and magnetizing force, respectively, at the knee of the magnetization curve, H_0 = mean value of magnetizing force, B_r = mean value of magnetic flux density, H_p = fictitious magnetizing force corresponding to B_r.*

an unknown auxiliary quantity used to define the unknown parameter α, see Fig. 7, in the relation

$$\cos\alpha = \frac{B_p - B_k}{B_r\sqrt{2}} \quad \ldots \quad \ldots \quad \ldots \quad (5)$$

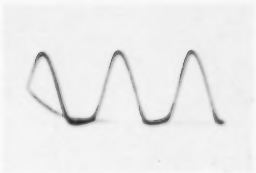
in which B_k denotes the magnetic-flux density at the knee of the magnetization curve. We have then with the notation of Fig. 7 and with equations (2) and (5) for the instantaneous value of magnetizing force the expressions

$$h \Big|_{0}^{\pi-\alpha} = H_k + \frac{b - B_k}{\mu_2\mu_0} = \frac{B_k}{\mu_1\mu_0} + \frac{B_p + B_r\sqrt{2}\cos\omega t - B_k}{\mu_2\mu_0} = \frac{1}{\mu_1\mu_0}\left[B_k + \frac{\mu_1}{\mu_2}B_r\sqrt{2}(\cos\alpha + \cos\omega t)\right] \quad (6)$$

and

$$h \Big|_{\pi-\alpha}^{\pi+\alpha} = H_k - \frac{B_k - b}{\mu_1\mu_0} = \frac{B_k}{\mu_1\mu_0} + \frac{B_p + B_r\sqrt{2}\cos\omega t - B_k}{\mu_1\mu_0} = \frac{1}{\mu_1\mu_0}\left[B_k + B_r\sqrt{2}(\cos\alpha + \cos\omega t)\right] \quad (7)$$

Fig. 8. *Oscillogram of the magnetizing force in a transductor core with sinusoidal transductor voltage and direct-current bias-magnetization.*



where $\mu_0 = 1.257 \times 10^{-6}$ henry/metre denotes the absolute permeability of free space expressed in the rationalized form of the m.k.s. system of units, and $\mu_1 = (\tan\theta_1)/\mu_0$ and $\mu_2 = (\tan\theta_2)/\mu_0$, see Fig. 7, are the relative incremental permeabilities of the high- and low-permeability regions, respectively, of the magnetization curve. The oscillogram of Fig. 8 verifies the typical wave form of h shown in Fig. 7. The Fourier analysis of h gives the series expression

$$h = H_0 + \sum_{n=1}^{\infty} H_n\sqrt{2}\cos n\,\omega t \quad \ldots \quad (8)$$

with the coefficients

$$H_0 = \frac{1}{\mu_2\mu_0}\left[B_r\sqrt{2}\left(\frac{\pi - \alpha}{\pi}\cos\alpha + \frac{\sin\alpha}{\pi}\right) + \frac{\mu_2}{\mu_1}\left(B_k + B_r\sqrt{2}\,\frac{\alpha\cos\alpha - \sin\alpha}{\pi}\right)\right] \quad (9)$$

for $n = 1$:

$$H_1\sqrt{2} = \frac{B_r\sqrt{2}}{\mu_2\mu_0}\left[1 - \frac{2\alpha - \sin 2\alpha}{2\pi}\left(1 - \frac{\mu_2}{\mu_1}\right)\right] \quad \ldots \quad \ldots \quad (10)$$

and for $n > 1$:

$$H_n\sqrt{2} = (-1)^n \frac{B_r\sqrt{2}}{\mu_2\mu_0} \left[\frac{2\sin n\alpha\cos\alpha - 2n\cos n\alpha\sin\alpha}{n(n^2-1)\pi}\right] \cdot \left(1 - \frac{\mu_2}{\mu_1}\right) \quad \ldots \quad (11)$$

It is the practice to measure I_L either with a r.m.s. instrument, $I_{L\,rms}$, or with a rectifier instrument, $I_{L\,rect}$. We obtain with equations (10) and (11) in the first case

$$I_{L\,rms} = k_i \frac{l}{N_L}\sqrt{H_1^2 + H_3^2 + H_5^2 + \ldots} \quad (12)$$

and in the second case

$$I_{L\,rect} = k_i \frac{l}{N_L}\frac{2\sqrt{2}}{\pi}(H_1 - \tfrac{1}{3}H_3 + \tfrac{1}{5}H_5 - + \ldots) \quad \ldots \quad (13)$$

where for Fig. 2(a): $k_i = 2$ ... (14a)

and for Fig. 2(b): $k_i = 1$ ... (14b)

N_L denotes the number of turns of a load coil, see Fig. 2.

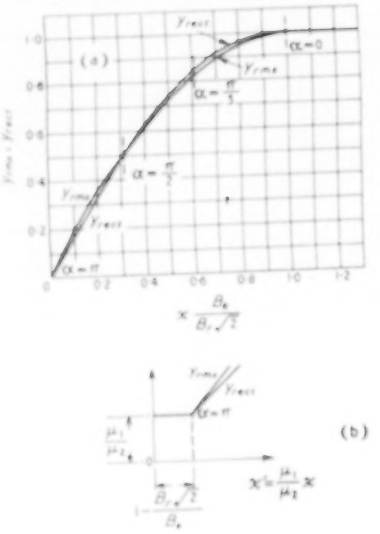


Fig. 9. *Calculated transductor characteristic for sinusoidal transductor voltage: (a) $\mu_1 \gg \mu_2$ (i.e., $\mu_2/\mu_1 = 0$). (b) Curves near the origin of the co-ordinate system for $\mu_2/\mu_1 \neq 0$.*

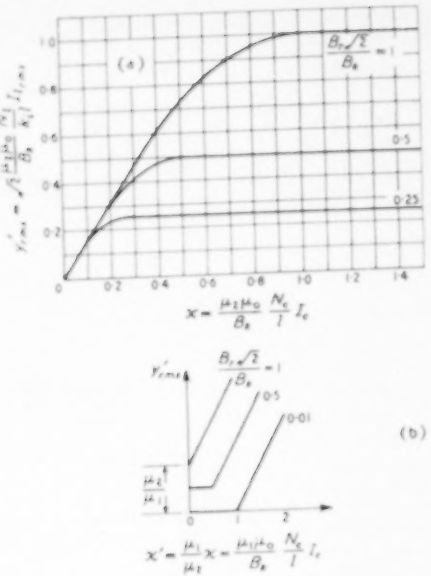


Fig. 10. *Family of calculated transductor characteristics for load current measured with a r.m.s. instrument: (a) $\mu_1 \gg \mu_2$ (i.e., $\mu_2/\mu_1 = 0$). (b) Curves near the origin of the co-ordinate system for $\mu_2/\mu_1 \neq 0$.*

In the transductor characteristic the load current, $I_{L\,rms}$ or $I_{L\,rect}$ is related to the control current I_c derived from equation (4):

$$I_c = \frac{l}{N_c}H_0 \quad \ldots \quad (15)$$

in which H_0 is substituted from equation (9). For generalized presentation of the transductor characteristic it is convenient to give the co-ordinates of the characteristic as dimensionless quantities. We write with equations (15) and (9):

$$x = \frac{\mu_2\mu_0}{B_s}\frac{N_c}{l}I_c = \frac{B_r\sqrt{2}}{B_s}\left(\frac{\pi - \alpha}{\pi}\cos\alpha + \frac{\sin\alpha}{\pi}\right) + \frac{\mu_2}{\mu_1}\left(1 + \frac{B_r\sqrt{2}}{B_s}\frac{2\alpha\cos\alpha - \sin\alpha}{\pi}\right) \quad \ldots \quad (16)$$

with equations (12), (10), (11) and (14):

$$y_{rms} = \frac{\mu_2\mu_0}{B_r}\frac{N_L}{k_i l}I_{L\,rms} = \sqrt{\left[\left\{1 - \frac{2\alpha - \sin 2\alpha}{2\pi}\left(1 - \frac{\mu_2}{\mu_1}\right)^2\right\} + \left(1 - \frac{\mu_2}{\mu_1}\right)^2 \sum_{n=3,5,7\ldots}\left\{\frac{2\sin n\alpha\cos\alpha - 2n\cos n\alpha\sin\alpha}{n(n^2-1)\pi}\right\}^2\right]} \quad (17)$$

and with equations (13), (10), (11) and (14):

$$y_{rect} = \frac{\mu_2\mu_0}{B_r}\frac{N_1}{k_c l}\frac{\pi}{2\sqrt{2}} I_{1,\,rect}$$

$$= 1 - \left(1 - \frac{\mu_2}{\mu_1}\right)\left[\frac{2x - \sin 2x}{2\pi} + (-1)^{\frac{n-1}{2}}\right]$$

Fig. 9(a) shows the calculated curves of y_{rms} and y_{rect} as functions of $x \cdot B_k/B_r\sqrt{2}$ for

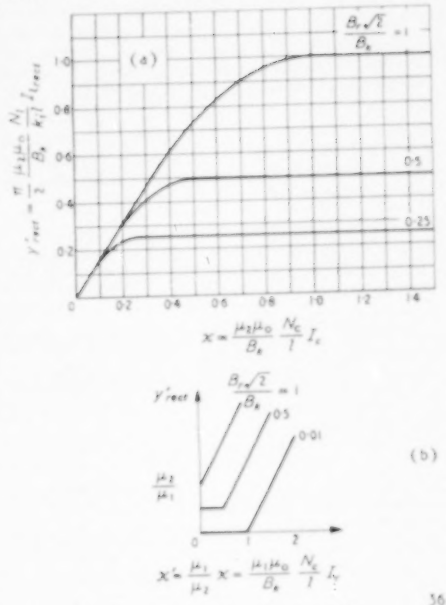


Fig. 11 (above). *Family of calculated transductor characteristics for load current measured with a rectifier instrument: (a) $\mu_1 \gg \mu_2$ (i.e., $\mu_2/\mu_1 = 0$), (b) Curves near the origin of the co-ordinate system for $\mu_2/\mu_1 \neq 0$.*

Fig. 12 (right). *Experimental transductor characteristics.*

$\mu_1 \gg \mu_2$; i.e., $\mu_2/\mu_1 = 0$ in equations (16), (17) and (18). The y_{rect} curve rises virtually linearly with x until $x = \pi/2$ (i.e., $x \cdot B_k/B_r\sqrt{2} = 0.32$), and then gradually bends to a constant value, $y_{rect} = 1$ for $x \cdot B_k/B_r\sqrt{2} \geq 1$. The y_{rms} curve is slightly above y_{rect} for $x > \pi/2$ and slightly below for $\pi/2 > x > 0$.

The curves for y_{rms} and y_{rect} in Fig. 9(b) are near the origin of the co-ordinate system and are calculated for $\mu_2/\mu_1 \neq 0$. For $0 \leq x' \leq (1 - B_r\sqrt{2}/B_k)$ we have $y_{rms} = y_{rect} = \mu_2/\mu_1$ = const. provided that $B_r\sqrt{2}/B_k < 1$. The incapacity to control the load current in this region is explained in terms of Fig. 7: direct-current

$$\sum_{n=3,5,7,\ldots} \left[\frac{2\sin nx\cos x - 2n\cos nx\sin x}{n^2(n^2-1)\pi}\right] \quad . . \quad (18)$$

bias-magnetization is small (i.e., $H_0 < H_k$), and the entire cycle of alternating magnetization is below the knee of the magnetization characteristic; i.e., $h < H_k$ throughout a magnetization cycle.

It is desirable for practical evaluation to have the transductor characteristic presented with $B_r\sqrt{2}/B_k$ as parameter. We therefore introduce, with equations (17) and (18),

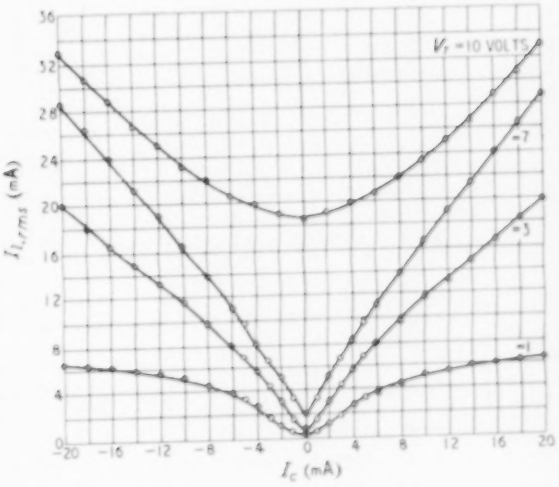
$$y'_{rms} = \frac{B_r\sqrt{2}}{B_k} y_{rms}$$

$$= \sqrt{2}\frac{\mu_2\mu_0}{B_k}\frac{N_1}{k_c l} I_{1,\,rms} \quad . . \quad . . \quad (19)$$

and $y'_{rect} = \frac{B_r\sqrt{2}}{B_k} y_{rect}$

$$= \frac{\pi}{2}\frac{\mu_2\mu_0}{B_k}\frac{N_1}{k_c l} I_{1,\,rect} \quad . . \quad . . \quad (20)$$

The families of characteristics shown in Figs. 10(a) and 11(a) represent y'_{rms} and y'_{rect}, respectively, plotted as functions of x, see equation (16), for $\mu_1 \gg \mu_2$ (i.e., $\mu_2/\mu_1 = 0$), and Figs. 10(b) and 11(b) give y'_{rms} and y'_{rect}, respectively, near the origin of the co-ordinate system for $\mu_2/\mu_1 \neq 0$. It is interesting to note from the curves of Figs. 10(a) and 11(a) that the condition $B_r\sqrt{2}/B_k = 1$ gives the largest range of 'linear' control. Figs. 10(b) and 11(b) further show that there is no 'backlash' of control at zero control



current when $B_r\sqrt{2}/B_k = 1$. The initial load current, at $I_c = 0$, can be made small by having a core material with a high value of μ_1.

The characteristics of Fig. 10 are verified by the experimental curves of Fig. 12 obtained with the transductor voltage V_T as parameter the control current I_c having positive and negative polarity. The curve for $V_T = 7$ volts corresponds to $B_r\sqrt{2}/B_k = 1$ and shows a sharp-cornered V-form. The curve for $V_T = 1$ volt is rounded at the bottom, short in the linear range, and bends to a constant value as predicted by the general form of the curves of Figs. 10(a) and 10(b). In the case of $V_T = 10$ volts we have a magnetization loop stretching into both regions of magnetic saturation (see, e.g., Fig. 4a) with a consequent high value of load current when $I_c = 0$.

Transductor Control Sensitivity

Let ΔI_c be a small change of control current and $\Delta I_{L\,reset}$ the corresponding change of load current. The transductor control sensitivity, S, is then defined by the relation

$$S = \frac{\Delta I_{L\,reset}}{\Delta I_c} \qquad \ldots \qquad (21)$$

or, with equations (16) and (18),

$$S = \frac{2\sqrt{2}}{\pi} \frac{k_l l}{N_l} \frac{B_r}{B_k} \frac{\Delta y_{reset}}{\mu_1 \mu_0 \Delta x} \qquad \ldots \qquad (22)$$

$$N_c \mu_1 \mu_0$$

On the linear part of the y_{reset} curve, Fig. 9(a), and for $\mu_1 \gg \mu_2$ (i.e., $\mu_2/\mu_1 = 0$), we have with equations (16) and (18)

$$\frac{\Delta y_{reset}}{\Delta x} = \frac{y_{reset}}{x} \text{ at } x = \pi/2$$

$$= \frac{\pi}{2} \frac{B_k}{B_r\sqrt{2}} \qquad \ldots \qquad \ldots \qquad \ldots \qquad (23)$$

Equation (23) substituted in (22) gives, with equation (14),

$$S = k_l \frac{N_c}{N_l} \qquad \ldots \qquad \ldots \qquad \ldots \qquad \ldots \qquad (24)$$

This means that in the case of $\mu_1 \gg \mu_2$ the current control sensitivity on the linear part of the transductor characteristic is determined only by the turns ratio between control winding and load winding, apart from the factor $k = 2$ if the transductor is of the parallel type.

Conclusion

The transductor control characteristic is determined by the shape of the magnetization curve of the transductor core and by the ratio between the peak value of alternating magnetic-flux density and the value of magnetic-flux density at the knee of the magnetization curve. The control characteristic has an optimum form in regard to range of linear control, low initial load current and absence of backlash of control at zero control current if the magnetization curve has a sharp bend at the knee and a high value of initial permeability, and when the peak value of alternating magnetic-flux density equals the value of magnetic flux density at the knee of the magnetization curve.

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TRANSMISSION FACTOR OF DIFFERENTIAL AMPLIFIERS

By D. H. Parnum, A.R.C.S., Ph.D., F.Inst.P.

SUMMARY.—It is shown that the quantity 'discrimination factor' (anti-phase/in-phase gain ratio) is insufficient to define the performance of a differential amplifier, and that a quite different quantity, the 'transmission factor,' sets the limit to the interference/signal ratio that can be tolerated. The likely values of this quantity are worked out for typical input stages and found to be insufficient for many purposes; an adjustable control is described which is capable of giving a very high value. Further analysis shows that the discrimination and transmission factors of all stages must be considered to obtain the overall transmission factor for a complete amplifier, and this leads to recommendations for the design of the first two stages.

AN excellent paper by Johnston[1] appears to show confusion on a highly practical point which has seldom been discussed by other writers. It concerns the so-called 'discrimination factor' of balanced or differential amplifiers. For clarity a brief recapitulation of the function and principles of these amplifiers will first be given.

The Coupled-cathode Input Stage

Fig. 1 shows a differential input stage of the coupled-cathode type. An 'anti-phase' (or 'differential') signal is one applied between the two grids; the signal currents cancel in R_c, there is no negative feedback, and an output appears at full gain between the anodes. An 'in-phase' signal is one applied between earth and the two grids in common; negative feedback occurs in R_c and the anode output is much less than in the first case. If the two halves of the stage are identical, moreover, the in-phase outputs are equal and the voltage difference between the anodes is zero; i.e., there is no anti-phase output.

This arrangement is used chiefly in medical or biological applications, where the input grids are connected to electrodes placed on (or in) the subject. The small voltages generated by the subject, which it is required to measure, are then applied between the grids as an anti-phase signal. Generally, however, there is stray capacitance coupling to the electrodes from an interfering source (most commonly 50 c/s), which gives rise to unwanted in-phase signals. The *balanced* property of the amplifier (i.e., equality of the two halves) means that these signals will not cause an anti-phase output and so mask the wanted signals. The difference in anti-phase and in-phase *gain* is required for a different reason. Interference signals are commonly many times (100—1,000) the full drive wanted voltage, and would hopelessly overload the later stages if amplified at full gain. These two properties, of balance and gain ratio, are therefore different in nature and required for different purposes.

Many circuits have been devised for securing a high gain ratio, that of Fig. 1 being one of the simplest and most effective. For this stage Johnston quotes the well-known formulae:

Anti-phase gain, $M = \mu R/(r_a + R)$, . . (1)

In-phase gain, $\bar{M} = \mu R/\{r_a + R + 2R_c(\mu + 1)\}$ . . . . . (2)

He then calls the ratio $M/\bar{M}$ the discrimination factor, F:

$$F = \{r_a + R + 2R_c(\mu + 1)\}/(r_a + R)^* \qquad (3)$$

In what follows it will be convenient to take average figures for μ, r_a and R; the numerical results are not vastly different over the range of triodes available. We shall take $\mu = 40$, $r_a = 20$ kΩ, $R = 100$ kΩ.

It is seen from (3) that F approaches infinity as R_c is made very large. If R_c is a resistor, it cannot be increased indefinitely without causing

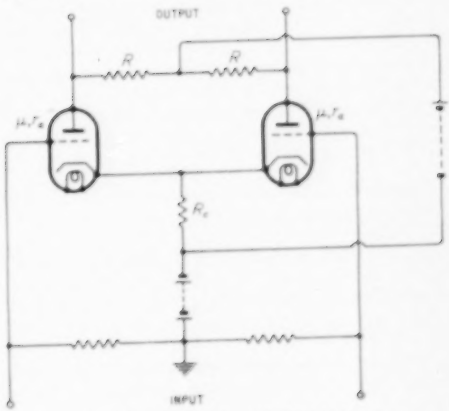

Fig. 1. *Cathode-coupled input stage.*

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* In Johnston's paper the denominator of this expression is misprinted.

an excessive static voltage drop; the value of R_s seldom exceeds $0.5R$. We then find that $F = 35$. This is not very large, and Johnston describes an ingenious method in which R_s is the slope resistance of a pentode. In this way a value of 5 MΩ can be obtained with a static drop of only 70 V; and F becomes 3,400.

Discrimination and Transmission Factors

Johnston's paper suggests that the attainment of high F is all that is needed to eliminate interference troubles. Thus, he criticises the low value obtained with a resistor for R_s, and advocates the pentode as a means of raising F to a satisfactory value. The implication is that if, say, $F = 3{,}000$, interference can be tolerated which is 3,000 times the minimum wanted signal. This is not the case. The point which Johnston does not mention is that *an in-phase input can produce an anti-phase output*. He does not, in fact, consider the previously-mentioned distinction between the 'balance' and 'gain ratio' properties. His formulae (1)—(3) involve mean parameters and so cannot take account of inequalities between the two halves of the stage. It is these inequalities, which are inevitable in a stage that has no adjustable control, which cause the spurious output to occur. Once it has occurred no discrimination or rejection process in the rest of the amplifier can remove it, because it is an anti-phase voltage and is indistinguishable from a genuine signal.

Before considering the analysis of a stage with unequal halves, as shown in the equivalent circuit of Fig. 2, it is useful to look at the matter physically and see how the anti-phase output can arise. It is obvious that if $R_s = 0$, the two halves function without interaction; they amplify the equal input voltages slightly differently, and the difference at the anodes is an anti-phase voltage. But if R_s approaches infinity, so that the in-phase output is zero, the situation is at first sight not the same. No in-phase current can flow in an infinite R_s; hence no such current can flow in R_1 and R_2. From this it is an easy, but false, step to the conclusion that no signal current of any kind flows in R_1 and R_2, so that there is no output voltage of any kind. The fallacy of this reasoning is that signal current can circulate round R_1, R_2, and the valves without flowing through R_s at all, and in fact it will do so if the halves are not identical. This can be seen from Fig. 2 by considering that μ_1 and μ_2 are not equal; the fictitious voltages in the two valves are then unequal and the resultant voltage must act round the closed circuit of R_1, R_2 and the valves. The current thus driven round the circuit gives rise to a pure anti-phase output. The mean value of this output (with respect to earth) is zero, which agrees with the statement that there is no in-phase output.

The term 'transmission factor' is suggested for this property of a stage (or complete amplifier) and is defined in the following way. Suppose that an in-phase input e gives rise to an anti-phase output Ke. The anti-phase input which would produce the same output is Ke/M. An in-phase signal therefore has to be M/K times as large as an anti-phase to produce the same effect, and this ratio is defined as the transmission factor, H:

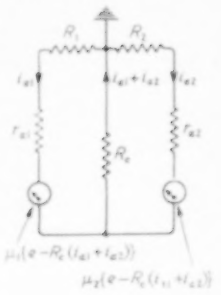

Fig. 2. *Equivalent circuit of Fig. 1 with unequal halves. In-phase input between each grid and earth = e.*

$$H = M/K \qquad (4)$$

The name 'transmission factor' has been proposed because 'discrimination factor' is by now probably firmly attached to the anti-phase/in-phase gain ratio. It is not ideal, but suggests to the writer, at least, the concept of the stage 'transmitting' a fraction of an in-phase voltage as anti-phase. If an amplifier is perfectly balanced, an in-phase voltage applied at the start does not appear in the final output at all; i.e., it is simply not transmitted. In an article by the writer[2] the term 'output ratio' (to contrast with 'gain ratio') is used; but this term now seems to be insufficiently descriptive.

Turning now to the analysis of Fig. 2, the result for K is:

$$K = \frac{\mu_1 R_1 r_{a2} - \mu_2 R_2 r_{a1} + (\mu_1 - \mu_2)\{R_1 R_2 + (R_1 + R_2)R_s\}}{(r_a + R)\{r_a + R + 2R_s(\mu + 1)\}} \qquad (5)$$

The numerator of this expression is accurate; the denominator is symmetrical in R_1, R_2, etc., and these quantities have been replaced by a common value R, etc.

To cast (5) into a useful form we put

$R_2 = R$, $R_1 = R + \Delta R$, etc.

It is then found that M is a factor of K, and we finally have from (4) the approximate value

$$H = \frac{r_a + R + 2R_s(\mu + 1)}{r_a} \Big/ \left[\frac{\Delta R}{R} - \frac{\Delta r_a}{r} + \frac{\Delta \mu}{\mu}\left(\frac{r_a + R + 2R_s}{r_a}\right)\right] \qquad (6)$$

The approximation is that $\Delta\mu/\mu$ should also be multiplied by $(1 + \Delta R/R)$; this gives rise to a second-order product. It is shown later that the $\Delta\mu/\mu$ term is nearly always predominant, and in that case the second-order product causes only a small change in this predominant term.

A number of useful deductions can be made from (6) by considering numbers and certain special cases. Assumptions will be made about the likely values of $\Delta\mu/\mu$, etc.; those chosen are:

(i) $\Delta\mu/\mu$, $\Delta r_a/r_a$, and unintentional values of $\Delta R/R$ do not exceed 10%.

(ii) R/r_a is not less than 3.

Three special cases are of interest.

Case A. $R_c = 0$.

This will substantially be so when the stage has a bias resistor only; i.e., no attempt is made at discrimination.

It may happen that the square bracket in (6) is zero, so that $H = \infty$ and the stage is perfect; but we are concerned with the lower limit of H in the worst case. This occurs when $\Delta\mu/\mu$, etc., are all numerically equal but such that they all add up. The square bracket then becomes

$$\frac{3r_a + R}{r_a} \cdot \frac{\mu}{\Delta\mu},$$

and we get for H:

$$H = \frac{r_a + R}{3r_a + R} \cdot \frac{\mu}{\Delta\mu}.$$

On the assumption made about R/r_a, the coefficient of $\mu/\Delta\mu$ lies between 2/3 and 1. As we only want to know the order of H in the worst case, it is convenient and sufficient to say that approximately

$$H = \mu/\Delta\mu.$$

On the assumption made about $\Delta\mu/\mu$, this may be as low as 10.

Case B. $R_c = 0.5R$.

This value is not likely to be exceeded when R_c is a resistor.

Again taking the most unfavourable combination in the bracket, we get

$$H = \frac{\mu + 2 + r_a/R}{2 + 3r_a/R} \cdot \frac{\mu}{\Delta\mu}.$$

The numerator of the coefficient is effectively μ; the denominator lies between 2 and 3. Following the course taken for Case A, it is again sufficient to say that approximately

$$H = \tfrac{1}{2}\,\mu^2/\Delta\mu.$$

This is a considerable improvement on Case A; for $\mu = 40$, H is 20 times as large. The likelihood of a fortuitous high value is decreased, however. The coefficient of $\Delta\mu/\mu$ in (6) is now not less than 7, so that unless $\Delta\mu/\mu$ is less than 1/7 of the other inequalities the term involving it will predominate. A study of (6) then shows that the lower limit deduced for H becomes the actual value. At the best, then, H may not exceed 200.

Case C. $R_c = \infty$.

This is substantially approached when a pentode is used for R_c.

It is at once seen from (6) that, without any approximations,

$$H = \mu^2/\Delta\mu.$$

This is only twice as good as Case B, and is an improvement which hardly justifies the trouble of using a pentode; other reasons, however, are given below which necessitate its use in the first stage. The value of H may be no higher than 400; if this is compared with the corresponding figure for F of 3,400, it is seen what a false idea F can give of the performance of the stage. It should also be noted that, when R_c is really infinite, we have a case where the in-phase gain is zero, but there is still a finite anti-phase transmission given by the above value of H.

As a tailpiece to this section, the following paragraph from Johnston may be quoted:

"The discrimination factor is increased in proportion to the cathode load. If adjustments are made to the inequalities of characteristics in a pair of valves with differential input and output, it is possible to realise a factor of 100,000 : 1; using components of commercial tolerance 3,000 : 1 is obtained in the circuit of Fig. 10(a) without further adjustment."

This paragraph appears to show complete confusion between discrimination and transmission factors. "Adjustments to inequalities" cannot alter F (which depends on mean parameters only), but can alter H, and the figure of 100,000 : 1 must be what we have defined as H. The figure of 3,000 : 1 is then stated to refer to the same quantity, so that it must also be H; yet it is really F. It is impossible to believe that Johnston is unaware of the distinction, but the writer can find no interpretation of the quoted paragraph which is self-consistent.

Attainment of High Transmission Factor

The values of H required are commonly very high. In a practical case of the writer's experience, the observed interference level (across 1-MΩ grid resistors) was 500 mV, while the minimum desired signal was 10 μV; thus H had to be at least 50,000. None of the above values approach this. It is, therefore, natural to enquire whether an adjustable control can be added to the stage to raise H to any given value. It is seen from (6) that the problem is to make the square bracket approach zero.

Control of ΔR by making the anode loads

variable is tempting; it has been used by some writers with possibly this end in mind, but examination of (6) shows it to be generally impracticable. When $R_a = 0.5R$ the $\Delta\mu/\mu$ term is nearly 1, and one anode load would have to be nearly zero to compensate for it; when $R_a = 5$ MΩ the term is 50 and control by ΔR is impossible. The same remarks apply to Δr_a. It is apparent from these examples that the $\Delta\mu/\mu$ term is dominant and the correct course is to vary $\Delta\mu$.

Fortunately, this can be done very easily by the circuit of Fig. 3. Each valve has a variable shunt from anode to cathode. By Thévenin's theorem, when a valve of parameters μ, r_a is shunted by S it behaves like one of parameters $\mu S/(S + r_a)$ and $r_a S/(S + r_a)$. The variation of $\Delta\mu$ is thus produced; the associated variation of Δr_a is, as just explained, of little effect in comparison.

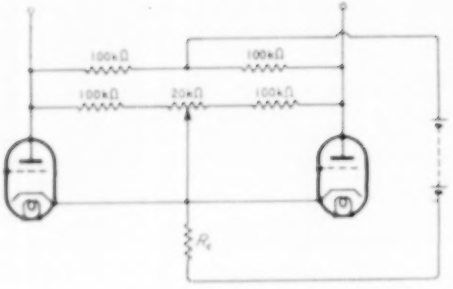

Fig. 3. *Elements of shunt control circuit to raise value of H.*

The values shown in Fig. 3 have been used with an ECC35. For this valve r_a is about 30 kΩ; the reduction of μ from one end of the control to the other therefore ranges from 77% to 92%. The mean is 74.5%; thus the total range of variation is 18% of the mean, and the circuit will cover differences of this amount. There is some loss of anti-phase gain, about 20%; this is a small price to pay for the advantages gained. The valve is working into a relatively low anode load, but this is immaterial as the anti-phase signal level is far too small to cause distortion.

One practical point which must be mentioned is that the shunt circuit upsets the static conditions, and allowance must be made for this when setting-up the stage. Theoretically a blocking capacitor avoids this trouble; but the compensation then fails progressively as the frequency is lowered, and it is impossible to maintain a large H, even at 50 c/s, without an impracticably large capacitance.

On test the circuit has given very satisfactory results. A value of $H = 50{,}000$ was easily attained. The sensitivity of the detecting circuit following the stage did not permit any greater value of H to be measured, and no attempt was made to set up a special circuit for doing so. The balance point did not appear to drift with time. It is probably sensitive to supply voltage; but this is unimportant because the supplies to a first stage of this nature must always be stabilized for other reasons.

The control naturally requires re-adjustment when the valve is changed. It is, however, independent of the input circuit conditions, and is therefore suitable for any application.

Overall Transmission Factor

The attainment of a large H in the first stage is a necessary condition for success, but not a sufficient one. Even if H is made infinite by careful adjustment, the stage still has a finite in-phase gain, and will pass some in-phase voltage. This will be applied to stage 2; if H for this stage is finite, it will give rise to an anti-phase output. Similarly stage 2 will pass some in-phase voltage to stage 3, which will in turn give rise to an anti-phase output, and so on. Clearly the overall effect depends on the transmission and discrimination factors of each stage. In general, the anti-phase output from one particular stage will predominate, and the design problem is to ensure that this one does not exceed the permitted amount.

The way in which the overall transmission factor H is built up is as follows. We shall write:

Anti-phase gain of stage n: M_n.

Discrimination factor of stage n: F_n.

Transmission factor of stage n: H_n.

It follows that:

In-phase gain of stage $n = M_n/F_n$.

Considering an in-phase input of 1 V to stage 1, we have the conditions of Table 1.

The last column shows the equivalent anti-phase voltages at the input which would give rise to the actual anti-phase transmission from each stage. We can therefore add all these terms to give the total equivalent input which would produce the total effect in the whole amplifier; this total input is by definition $1/H$. Thus

$$1/H = 1/H_1 + 1/F_1H_2 + 1/F_1F_2H_3 + \cdots \quad (7)$$

Exact summation of this expression is unimportant. It is only necessary to examine the denominators to ensure that the smallest is sufficiently large, and that none of the others is less than about twice the smallest. To continue with our example for which H was required to be 50,000, suppose we make $H_1 = 50{,}000$ by

adjustment of the control in the first stage; then F_1H_2 must be not less than about 100,000. If stage 1 has a resistor for R_c, F_1 will be about 30, so that H_2 must be about 3,000. This is not possible unless stage 2 has specially selected valves or an adjustable control, and either of these courses is inconvenient. But if R_c is a pentode, F_1 will be about 3,000, and H_2 need be no more than 30. This cannot be guaranteed if stage 2 has no cathode resistor, but can be if it has one of no more than about $R/5$, which is about the lowest value likely to be used. Thus the combination which gives a satisfactory result and is the least trouble is: pentode and shunt control for stage 1, resistor for stage 2.

The remaining terms in (7) are now automatically large enough. For $R_c = R/5$ we get $F_2 = 13$, so that F_1F_2 is about 40,000. This occurs as a factor in all the remaining terms; they also each contain an H which cannot be less than 10, even with no cathode resistor; thus they cannot be less than 400,000.

The question of overloading by in-phase interference is also settled. The maximum expected interference is 50,000 times the minimum signal; but the overall in-phase gain is at the most $1/F_1F_2 = 1/40{,}000$ of the anti-phase. Thus the total in-phase voltage arriving at the output stage can be very little more than the *minimum* signal, so that overloading or cross-modulation is utterly impossible.

It should be mentioned finally that when the shunt control is fitted to the first stage, it is possible to obtain a high overall H without the above precautions. By slight misadjustment of the control, the anti-phase transmissions of the various stages can obviously be made to cancel. Such a condition, however, is likely to be more sensitive to drift, and the control will require re-setting when *any* of the valves are changed. The proposed arrangement is sensitive to conditions in the first stage only.

Conclusions

Without an adjustable control in the first stage of a balanced amplifier, in-phase interference will in general produce an intolerable output which appears to be due to a spurious signal. The use of a cathode pentode is insufficient to prevent this, and may result in a transmission factor of no more than 400. A successful control has been tested which permits H to be at least 50,000. Further consideration shows that it is still necessary to use a pentode in the first stage, and a cathode resistor in the second, if this value of H is not to be dominated by the effects of the second and later stages. If these precautions are adopted, the remaining stages need no special attention, and overloading by in-phase interference is impossible.

TABLE 1

	In-phase input	In-phase output	Anti-phase output	Gain of preceding stages	Equiv. input to stage 1
Stage 1	1	M_1/F_1	$1/H_1$	1	$1/H_1$
Stage 2	M_1/F_1	M_1M_2/F_1F_2	M_1/F_1H_2	M_1	$1/F_1H_2$
Stage 3	M_1M_2/F_1F_2	$\frac{M_1M_2M_3}{F_1F_2F_3}$	$M_1M_2/F_1F_2H_3$	M_1M_2	$1/F_1F_2H_3$

Acknowledgment

My thanks are due to the Cambridge Instrument Company for permission to publish this paper and to make use of the experimental material.

REFERENCES

[5] D. L. Johnston, "Electro-Encephalograph Amplifier", *Wireless Engineer*, August 1947, Vol. 24, pp. 231–242; September 1947, Vol. 24, pp. 271–277; October 1947, Vol. 24, pp. 292–297.

[6] D. H. Parmee, "Biological Amplifiers", *Wireless World*, November 1945, Vol. 51, pp. 337–340; December 1945, Vol. 51, pp. 373–376.

ERRATA

In the article "Phase-Shift Oscillator," by W. C. Vaughan, in the December 1949 issue, two errors occurred. In Equ. (16), p. 393, the last term should be "6," not "1" and Equ. (18), p. 393, the denominator of the second fraction should be "$6 + n$," not "$4 + n$." The equations appear correctly in Table 1.

Two errors occurred in the letter from R. Fürth, "Spontaneous Fluctuations in Double-Cathode Valves," on p. 33 in the January issue. Equation (1) should read:

$$(\delta I)^2_f = 4\,kTg_s\,\beta\Delta f$$

and Equation (2) $g_s = \left(\frac{\partial I}{\partial V_s}\right)_{I_s}$

An error occurred in the advertisement of H. W. Sullivan on p. 16 of the March issue. The range of frequencies over which the power factor of this firm's variable air condensers is below 0.00001 was given as 10^3 to 10^8 c/s instead of 10^5 to 10^8 c/s.

CORRESPONDENCE

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Non-Linear Effects in Rectifier Modulators

SIR,—The non-linear operation of an ideal ring modulator was analysed in *Wireless Engineer*, Vol. 26, May 1949, p. 177, and it was stated that similar results hold for a Cowan modulator. This was criticized by D. G. Tucker and E. Jeynes in Vol. 27, February, 1950, p. 66, who raised various objections and presented many interesting new points. These finer points of the theory can only be examined on the basis of a more elaborate analysis taking into account the finite passing blocking conductances, g_r and g_s, of the rectifiers and the finite carrier source resistance r. The results of this analysis are outlined below for (1) a ring modulator, (2) a Cowan modulator with a complementary rectifier connected in opposite polarity, (3) a single Cowan modulator. In cases (1) and (2) the results obtained by making $g_r = \infty$ and $g_s = 0$ coincide with those of May 1949, whatever the carrier source resistance may be. On the other hand, in case (3) the previous result is only valid for $r = 0$, and, for finite values of r, the new effects mentioned by Tucker and Jeynes are quantitatively explained. Thus, there is no reason to question the validity of a theory based on a discontinuous rectifier law. Even if it were proved that the exact shape of the rectifier characteristic in the transition interval has an important role in intermodulation processes, it would be interesting to separate this effect from similar ones already produced with a rectifier characteristic nearer to the ideal.

In any case, in practical modulators, additional complications are caused by selective termination and unbalance effects.

The ring modulator is assumed to work between resistances R and the instantaneous signal e.m.f. is denoted by xv_s. If ε_c is the instantaneous carrier the carrier voltage in linear operation is given by

$$v_c = \varepsilon_c/f \; ; \; f = 1 + 2r\,(g_r + g_s) \quad \ldots \quad (1)$$

and the voltage v_c so defined will be considered as the reference variable. When overload occurs, the modulator may be unbalanced and, if r is not zero, the actual carrier voltage v_c' will differ from v_c and contain a signal component. A first overload point occurs when the signal voltage is slightly larger (due to the shunt effect of the other blocking rectifier) than the carrier voltage across one nominally blocking rectifier; i.e., for $v_c' = \pm\, av_s$, where

$$a = (1 + g_s R)^{-1} \quad \ldots \quad (2)$$

A second overload point occurs when the signal current is slightly larger than the carrier current in a nominally passing rectifier; i.e., for $v_c' = \pm\, bv_c$, where

$$b = (1 + g_r R)^{-1} \quad \ldots \quad (3)$$

The signal leak into the carrier source can be calculated and v_c', expressed in terms of v_s and v_c. This does not alter the first overload condition but changes the second one into $v_c = \pm\, b'v_c$, where

$$b' = bk/f \; ; \; k = \frac{1 + 2r\,(g_r + g_s + 2Rg_rg_{ms})}{1 + \frac{1}{2}\,(g_r + g_s)\,R} \quad \ldots \quad (4)$$

It is convenient to represent the condition of the rectifiers in the plane of co-ordinates v_s, v_c. This is done in Fig. 1 where $\pm\, a$ and $\pm\, b'$ denote the slopes v_c/v_s of the straight lines separating the various modulator configurations. Passing rectifiers are shown in full and blocking rectifiers in dotted lines. When v_c and v_s are harmonic functions with amplitudes V_c and V_s the point representing the instantaneous configuration of the modulator traces a Lissajous figure inscribed in a rectangle of co-ordinates $(\pm\, V_s, \pm\, V_c)$. The instantaneous output voltage is computed for each configuration by elementary network theory and all the voltages are combined into a single expression by using the discontinuous function of two variables*

$$F(x, y) = \tfrac{1}{4}x\,[\mathrm{sign}(y + x) + \mathrm{sign}(y - x)] + \tfrac{1}{4}y\,[\mathrm{sign}(y + x) - \mathrm{sign}(y - x)] \quad \ldots \quad (5)$$

studied in connection with the ideal modulator. The output voltage is

$$v = p[F(v_s, av_s) + qF(v_s, b'v_s)] \quad \ldots \quad (6)$$

where

$$p = \tfrac{1}{2}(g_s - g_r)\,Rf/z \quad \ldots \quad (7)$$

$$z = 1 + \tfrac{1}{2}(g_r + g_s)R + r(3g_r + g_s) + g_rgR\,(3g_r + g_s) \quad \ldots \quad (8)$$

$$q = (1 + 4rg_r)/k \quad \ldots \quad (9)$$

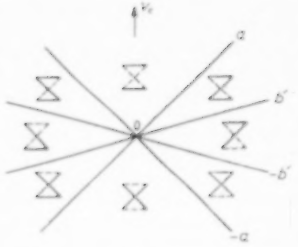

Fig. 1.

For an ideal modulator, $a = p = 1$, $b' = 0$ and the second overload occurs at infinite signal voltage; thus $v = F(v_s, v_c)$ as obtained previously. When v_c and v_s are harmonic functions, the double Fourier series expansion of (5) permits calculation of all intermodulation products. In particular, overload characteristics for the main sidebands are expressible in terms of elliptic integrals and can be computed numerically using, for instance, the tables of Jahnke-Emde, Dover, 4th Ed., p. 85. In practice, the second overload point occurs at very high signal voltages and the second term in (6) is reduced to its linear part $qb'v_c$. It can easily be checked that in linear operation the main sideband output takes the conventional value

$$V_{11} = \frac{2}{\pi}\,V_s k \; ; \; k = p(a + qb') = \frac{(g_r - g_s)\,R}{(1 + g_rR)\,(1 + g_sR)} \quad \ldots \quad (10)$$

Assuming that the matching condition $R = (g_sg_r)^{-\frac{1}{2}}$ holds and neglecting higher powers of the rectifier ratio $\alpha = (g_r/g_s)^{\frac{1}{2}}$, the intermodulation margin for small signal-to-carrier ratio is given by

$$20 \log \left|\frac{V_{11}}{V_{13}}\right| = 40 \log \left|\frac{V_c}{V_s}\right| + \left(\frac{3 + 8rg_r}{1 + 2rg_r}\right) 8.7 + 27.6 \text{ decibels} \quad \ldots \quad (11)$$

For $\alpha = 0$ (ideal modulator), this checks with the formula published previously and the additional dissipative

* Sign x = the signature of x and is defined as sign $x = +1$ when x is positive, sign $x = 0$ when $x = 0$, and sign $x = -1$ when x is negative.

term increases the intermodulation margin from 26x for $r = 0$ to 35x for $r = \infty$. For a matched carrier source $r = 1/2g_a$, and this term is 30.5x. The signal leaking into the carrier circuit has also been calculated: the voltage of the main products $f_c \pm f_s$ across the carrier source at impedance matching is approximately

$$3xV_s^3/8\pi V_c^2 ;$$

the leakage attenuation, taking into account the different signal and carrier impedances, is thus

$$40 \log |V_c/V_s| + 16 + 10 \log (1/x) \text{ decibels}$$

and is infinite for an ideal modulator.

The analysis of a Cowan modulator with a fifth rectifier in opposite connection leads to quite similar results if the signal impedances are denoted by $2R$ and the carrier source resistance by $2r$. The carrier voltage and e.m.f. are still related by (1) in linear operation. The modulator configurations are shown on Fig. 2. The first overload occurs as for the ring modulator but there are two further overload points characterized by the coefficient b defined in (3) and by

$$b'' = \frac{r(g_s - g_a)}{[1 + \frac{1}{2}R(g_s + g_a)]f} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (12)$$

As the configuration of Fig. 2 is not antisymmetrical with respect to V_s, intermodulation products of the type $mf_c \pm nf_s$ with m even or zero and n odd will also appear. These unimportant products complicate the discussion but are eliminated by considering only that part v', of the output voltage which is antisymmetrical with respect to v_s; i.e.,

$$v' = \tfrac{1}{2}[v(v_s) - v(v_s)].$$

This gives

$$v' = \tfrac{1}{2}p'[F(v_s, av_s) + (1 - z'/z)F(v_s, b''v_s) + \frac{z'}{z}F(v_s, bv_s)] \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (13)$$

where p and z have the signification given above and z' is deduced from z by interchanging g_s and g_a. For all practical purposes, only the linear part of the last two terms is of importance. This is

$$(1 - z'/z)\, b''v_s + z'bv_s/z = qb'v_s$$

and is identical to the linear part of the single second overload term of the ring modulator. Thus, with this approximation, (6) and (13) only differ by the factor $1/2$ (i.e., by 6 db) in all the products.

Using the same notation (impedances $2R$ and $2r$) for the single Cowan modulator, the positive carrier voltage in linear operation is

$$v_c = e_c/f' ; \quad f' = 1 + 2rg_s \quad \ldots \quad \ldots \quad \ldots \quad (14)$$

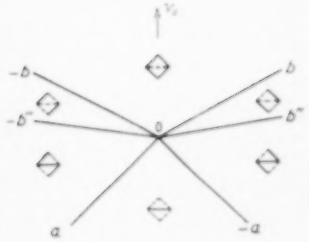

Fig. 2

The modulator configurations are defined in Fig. 3 with

$$c = \left(\frac{1 + 2g_s r'}{1 + 2g_a r'}\right) a \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (15)$$

This defines the first overload point which is strongly dependent on r. The second overload is of the same order as before but cannot *a priori* be neglected with respect to the first one. The antisymmetrical output voltage is

$$v' = \tfrac{1}{2}p''[F(v_s, cv_s) + F(v_s, bv_s)] \quad \ldots \quad \ldots \quad \ldots \quad (16)$$

with

$$p'' = \frac{\frac{1}{2}R(g_s - g_a)f'}{1 + \frac{1}{2}R(g_a + g_s) + r(g_a + g_s + 2Rg_sg_a)} \quad (17)$$

In linear operation this gives

$$V_{11} = \frac{1}{\pi}p'(c + b)\,V_s = \frac{1}{\pi}V_s k,$$

as expected, and the intermodulation ratio is

$$\frac{V_{13}}{V_{11}} = \frac{1}{24}\left(\frac{V_s}{V_c}\right)^2 \lambda \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (18)$$

for small signal, with

$$\lambda = \frac{c^3 + b^3}{c + b} = b^2[(c/b)^2 - c/b + 1] \quad \ldots \quad \ldots \quad (19)$$

As b is independent of r, this function will be a minimum for $c/b = 1/2$; i.e., for

$$r = \frac{\frac{1}{2} + R(g_s - 1\frac{1}{2}g_a)}{g_s - 2g_a - g_sg_aR} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (20)$$

which tends to R for good modulators. Assuming the usual matching condition $R = (g_sg_a)^{-1/2}$ and keeping only significant powers of $\alpha = (g_a/g_s)^{1/2}$, the function λ takes the following values for various values of r:

r	0	$1/2g_s$ (matching)	$R/2$ ($b = c$)	R (optimum)	∞
λ	1	$1/4$	α^2	$0.75\alpha^2$	α^2

V. Belevitch

Brussels.

Are Transit Angle Functions Fourier Transforms?

Sir,—Despite the time which has elapsed since the writer's letter on this subject appeared, no one has shown any signs of interest. As there are now many workers in the field, I feel that this is no accident, and that the difficulty remains unsolved. In this event it may be of some use to place on record the writer's further thoughts on the subject, particularly as the whole question of sound engineering of high-frequency devices is intimately bound up with the transit-angle functions associated with the various gaps or paths of their electrodes.

First, it may be said that transit-angle functions *are* Fourier Transforms, but that perturbation of the direct-current conditions by alternating forces renders it difficult to arrive at a true result. The fact that the Fourier Integral of a linear saw-tooth wave is convertible to the correct 'transit-time' solution is believed to be due to the fact that, when such a saw-tooth is perturbed in a certain way its Fourier Transform remains unchanged.

Secondly, in regard to the solution for, say, the distance travelled in a temperature-limited diode, it must be remembered that this is obtained as a result of a double integration. The double integral involved may in well-known manner be converted to a single integral, of the form

$$\sum_{n=-\infty}^{n=+\infty} \int_{t_0}^{t} (t-u)\, e^{jn\omega u}\, du \qquad \ldots \qquad (1)$$

where $n = 0$ gives the d.c. term and $n = 1$ the fundamental. Putting $n = 0$, we see that we only get a 'parabolic' pulse when the limits are inserted after integration; and this 'pulse' is in terms of $t - t_0$, not of $t' - t_0$ where $t'(= u)$ is the intermediate value of t. When $n = 1$, the integral for this term may be transformed to

$$e^{jn\omega t} \int_{0}^{\tau} u\, e^{-jn\omega u}\, du = \frac{x^2}{2(j\omega)^2} \cdot Y_2(x) \cdot e^{jn\omega t} \qquad \ldots \qquad (2)$$

Where τ is the 'unelapsed' part of the transit time, and $x = j\omega\tau$. This correct solution can be obtained, if one pleases, by regarding (1) as the 'incomplete' Fourier Transform of the unelapsed transit time $t - t'$.

I wish I were able to report success of a similar nature with the space-charge limited diode. A Fourier solution for the electron velocity has the same form as (2), but whereas this is correct, subject to allowance subsequently being made for 'variation time,' the temperature-limited diode velocity is already correct (including variation time effects), so this is not good enough. The reason why (2) is also correct is that when x is placed equal to a constant the variation time comes from the two equations

$$v = Y \frac{\tau^2}{2} [1 + \varepsilon_1 Y_3(x)]$$

$$\tau = Y \frac{\tau_0^2}{2}$$

(in other words $\tau_1 \tau_0 = -\frac{1}{3} \tau_0^2 x Y_3(x)$), so (2) is thus, in effect a solution for variation time; only v need be corrected, and, in the space-charge case, dv/dt. This brings us to the discrepancy in the space-charge case for the acceleration, and for the electric field, obtained as a first integration of the equation

$$d^2x/dt^2 = \frac{d^2v}{dt^2} = C \left(1 + \sum_{1}^{\infty} M_n e^{jn\omega t}\right)$$

where C is proportional to the space current per cm² of plane cathode. Transit-time method (a single integration, and subsequent 'variation time' correction applied) gives (for the first 2 terms)

$$\frac{dv}{dt} = C \left[\tau_0 + M_1 \frac{1 - Y_3(x)}{j\omega} e^{jn\omega t}\right]$$

where

$$\tfrac{1}{3}(1 - Y_3(x)) = x(1 - \tfrac{1}{2}x + \tfrac{1}{15}x^2 - \tfrac{1}{45}x^3 + \tfrac{1}{210}x^4 - \ldots)$$

while the Fourier Transform of a saw-tooth pulse (proportional to τ^2) gives a solution with the function

$$\frac{1}{2!\omega} x Y_2(x), \text{ where}$$

$$x Y_2(x) = x(1 - \tfrac{1}{3}x + \tfrac{1}{15}x^2 - \tfrac{4}{315}x^3 + \tfrac{1}{210}x^4 - \ldots)$$

exactly as for the velocity in the temperature-limited case. The differences are far greater than can be gauged by mere inspection of the power series expansions, moreover the Fourier method as above applied fails since the first term $\frac{1}{3}x$ is here wrong—it has to be $\frac{2x}{3}$, as is given by $1 - Y_3(x)$, in order to satisfy 'low frequency' conditions. If, however, it is noted that transit-time dynamical methods are as a rule as much simpler than the manipulation of Fourier Integrals, the engineer need have no qualms about using the less classical procedure.

As an example of great simplification by using 'transit time' may be mentioned the classical theory of radiation-collisions of electrons with atoms producing electromagnetic pulses as a result of sudden deceleration. The problem, dealt with by Lorentz, Larmor and Thomson[2] comes out much more simply if Fourier Integrals are dispensed with entirely, at any rate, in case $t_1 \ll t_2$. The connection between transit-time effects and electromagnetic radiation is, thus, incidentally, quite close—you cannot have the one without the other.

Fordcombe, Kent. W. E. BENHAM.

[1] *Wireless Engineer*, June 1949, p. 210.

[2] "Corpuscular Theory of Matter," by J. J. Thomson, 1907, p. 92. Constable.

Gain of Aerial Systems

SIR,—The point which I wished to make clear in my letter (*Wireless Engineer*, December 1949,) was that the end-fire theorem, stated by Mr. Bell in his paper in the September issue, did not apply to apertures in the diffraction sense. Practical cases of apertures in this sense are the mouths of mirrors and electromagnetic horns, and the outer surfaces of lenses. The argument advanced in my previous letter referred entirely to apertures in this sense, and not as Mr. Bell has implied in his reply (*Wireless Engineer*, January 1950), to arrays of conducting elements. As far as such arrays are concerned, I am in agreement with Mr. Bell that the gain may be doubled by using the array to produce an end-fire beam instead of a broadside beam. In view of the confusion which has arisen, it is worth while considering in what respects apertures of the diffraction type differ from arrays of conducting elements.

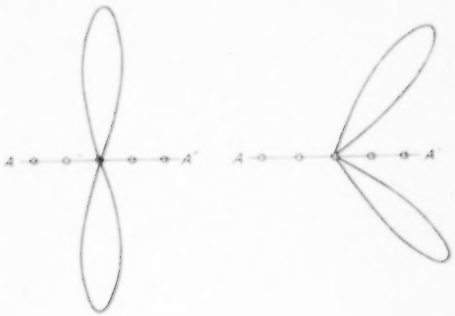

Fig. 1 Fig. 2

When the currents in the conducting elements of an array are all of the same phase, two beams will be radiated in opposite directions as shown in Fig. 1. Now, if the phases of the currents are suitably altered, it will be possible to swing each beam round towards the direction of the axis AA' as in Fig. 2. Eventually, a position will be reached when these two beams will overlap completely as in Fig. 3. The array is now acting as an end-fire radiator, and in view of the fact that the *two* broadside beams have coalesced into the *one* end-fire beam, it is reasonable to expect the gain of the system to be increased, as predicted by the end-fire theorem.

When the phase across a diffraction aperture is constant, however, there will be only one radiated beam; this is immediately obvious for any system such as a mirror, a lens or a horn. There is therefore no possibility of doubling the gain by overlapping two beams. In

fact, as shown in my last letter, the gain of the end-fire beam will be twice that of the broadside beam, only if the radiating aperture is surrounded by an infinite conducting sheet.

Fig. 3

My objection to Mr. Bell's statement of the end-fire theorem is that it is so phrased as to appear valid for either of the above types of aerial system. The true position is that the theorem is valid for an array of conducting elements, but only for an aperture in the diffraction sense, if it is surrounded by an infinite conducting plane. In the second case, it is the presence of the conducting plane, rather than the fact the aperture is acting as an end-fire radiator, which leads to the doubling of the gain.

Radar Research and
Development Establishment.

J. Brown

Phase-Shift Calculation

Sir,—In a recent letter (*Wireless Engineer*, January 1950) W. Saraga and J. Freeman describe a rapid, approximate, calculation of image phase-shift in multi-section Zobel filters. It is of interest to note that a more exact calculation can easily be carried out in the special case where all minima in the image attentuation are equal to each other.

Cauer has shown a method for determining the optimum distribution of the attenuation peaks in a low-pass filter where the equal minima of the image attenuation will fall in the frequency band $f' \leqslant f \leqslant \infty$ ($f' > f_2$, the cut-off frequency). The frequencies for the peaks are calculated by means of the elliptic sn-function. It is not very difficult to rearrange the optimal distribution to the band $f' \leqslant f \leqslant f'' < \infty$ with suitable frequency transformations. The m-values for a Zobel filter can then be calculated from the peak frequencies. Presumably this has been done by a great many involved in actual filter design.

A slightly different method also permits the phase-shift in the pass-band to be calculated very easily for the resulting filter, as well as the image attenuation for $f > f''$ or $f < f'$ and the effect of dissipation on filter performance.

First, the following frequency transformations of a well-known type are used :

$w = u + jv = \sqrt{1-(f_2/f)^2}$ for low-pass

$w = u + jv = \sqrt{1-(f/f_1)^2}$ for high-pass

$w = u + jv = \sqrt{\dfrac{f_2}{f_1}\,\dfrac{f_1^2-f^2}{f_2^2-f^2}}$ for band-pass, general case.

f_1, f_2 are the cut-off frequencies.

$w = u$ corresponds to an attenuation band

$w = jv$ corresponds to a pass band.

The frequency band $f' \leqslant f \leqslant f''$ is now transformed to $u' \leqslant u \leqslant u''$, where the image attenuation $A \geqslant A_{min}$ for N (>1) full-section peaks in the filter. By means of the rapidly convergent series-expansions for the Jacobi Theta-functions the following relations are established. (The infinite series can be terminated after only three terms or less with sufficient accuracy).

(1) Given : u', u'', N.

$$A_{min} = N \log_e \frac{1}{q} = \log_e z \text{ nepers}$$

$$q = \epsilon + 2\epsilon^5 \; ; \; \epsilon = \tfrac{1}{2}(1-\sqrt{u'/u''}) \,/\, (1+\sqrt{u'/u''})$$

(2) Given : A_{min}, N.

$$q = \exp\left[-(A_{min} + \log_e 2)/N\right]$$

$$\sqrt{u'/u''} = (1-2q+2q^4) \,/\, (1+2q+2q^4)$$

For the part of the attenuating range in question ($u' \leqslant u \leqslant u''$) we have :

$$A = A_{min} - \log_e |\cos Nz|$$

$$u = \sqrt{u'u''}(1-2q\cos 2z+2q^4\cos 4z) \,/\, (1+2q\cos 2z+2q^4\cos 4z)$$

where $0 \leqslant z \leqslant \pi$.

The attenuation peaks are placed at

$$u_n = u(z_n),\ z_n = (2n-1)\,\pi/2N,\ n = 1, 2, \ldots N$$

(For the low-pass and high-pass case u_n will be equal to the Zobel m-values).

By choosing a suitable value of z, corresponding values of u and A are readily calculated.

The image attenuation and phase-shift are now completely specified and can be determined in the ordinary way by adding the contributions from the separate filter sections. The following relations will, however, give easier calculations with little loss of accuracy.

In the *attenuating* band where $A < A_{min}$ [$u'' < u < \infty$ or $0 < u < u'$] we get :

$$A = A_{min} - \log_e\left\{\cosh\left[\frac{2y}{\pi}\,(A_{min} + \log_e 2)\right]\right\}$$

$(0 < y < \pi/2)$

$y = y_0 + 2\exp(\pi^2/\log_e q)\sin 2y_0$

$\cos y_0 = u/u'$ or $\cos y_0 = u''/u$ respectively.

[$\exp(\pi^2/\log_e q)$ is a small correction factor].

In the *pass* band ($0 \leqslant v \leqslant \infty$) the image phase-shift can be written as :

$$B = N(B_0 + q^2 \sin 2B_0)$$

where B_0 is the image phase-shift for a full-section filter with an attenuation peak at $u = \sqrt{u'u''}$.

The following equation can be used for calculation of the dissipation losses in the pass band etc.

$$\omega\frac{dB}{d\omega} = \frac{1}{s}N(1+2q^2\cos 2B_0)\,|\sin B_0|$$

In the attenuation range the peaks will be rounded off to finite values A_n due to filter losses :

$$A_n = A_{min} + \log_e\left(\frac{1}{d_L+d_C}\cdot N(1-\frac{8sq\sin z_n}{4q^2\cos^2 z_n}\right)$$

d_L, d_C = the dissipation factor for the coils and capacitors respectively.

In the equations above we have

$s = \left|\dfrac{w}{f}\dfrac{df}{dw}\right|$ which can be expressed in several ways, for instance.

$$s = \frac{f_2-f_1}{f_2+f_1}\cdot|\cosh^2\Gamma_k|$$

$\Gamma_k = A_k + jB_k$ is the complex image attenuation for an ideal half-section constant k-filter.

$|\cosh^2\Gamma_k| = \cos^2 B_k$ in a pass band

$|\cosh^2\Gamma_k| = \sinh^2 A_k$ in an attenuation band.

Stockholm,
Sweden.

Nils Olof Johannesson

R.E.C.M.F. EXHIBITION

The 7th exhibition of components, materials, test gear and valves opens at Grosvenor House, Park Lane, London, W.1. on 17th April. It is to be open for three days from 10 a.m. to 6 p.m. and admission is restricted to holders of invitation cards from the Radio & Electronic Component Manufacturers' Federation, 22, Surrey St., London, W.C.2.

NEW BOOKS

The Mathematics of Circuit Analysis

By Ernst A. Guillemin. Pp. 590 + xiv. John Wiley & Sons, New York and Chapman & Hall, 37 Essex St., London, W.C.2. Price 60s.

About 20 years ago the staff of the Department of Electrical Engineering at the Massachusetts Institute of Technology undertook to revise its entire presentation of the basic principles of electrical engineering. This revision resulted in the publication in 1943 of three volumes entitled "Electric Circuits" (a first course in circuit analysis for electrical engineers), "Magnetic Circuits and Transformers" (a first course for power and communication engineers), and "Applied Electronics" (a first course in electronics, valves and associated circuits). The volume under review was intended to be part of the same series, but publication was delayed by the war. Basic mathematical processes and methods are discussed; the chapter headings are Determinants, Matrices, Linear Transformations, Quadratic Forms, Vector Analysis, Functions of a Complex Variable, and Fourier Series and Integrals.

The object of the book is to explain how these basic methods and processes work; applications are deliberately kept for the companion volumes mentioned above. It is intended to stimulate interest and convey ideas rather than to be rigorous, and to develop a background of general understanding upon which the student whose mathematical ability is above the average may later build more carefully. Its scope is much wider than that of a book designed merely to help the student pass examinations.

Under the present British conditions, the time is hardly ripe for so fundamental a revision of the structure of electrical engineering courses as that undertaken at the M.I.T. While this may make it difficult to use the book directly as a text-book in this country, it contains a valuable collection of reference material chosen as the result of Prof. Guillemin's long practical experience of teaching electrical engineers. It therefore deserves careful study and should be available in libraries.

J. W. H.

Television for Radiomen

By Edward M. Noll. Pp. 595 + xii, with 356 illustrations. Macmillan & Co., Ltd., St. Martin's Street, London, W.C. 2. Price 52s. 6d.

This is an American book and deals solely with television equipment designed for the American standards. Its utility in this country is thus strictly limited by this fact.

It covers all aspects of television receivers in an elementary manner and includes transmitting material only in just enough detail to give the reader a picture of the complete system.

The method of treatment is usually to take a section of a receiver and to discuss it in general terms. Various alternative circuits are illustrated and described in detail and then a number of typical circuits, as used in American receivers, are given as examples of the way in which they take practical form. On these last circuits values of components are given.

As an example, the chapter on "Sweep Systems" starts with the basic method of generating a saw-tooth wave by the alternate charge and discharge of a capacitor. Blocking-oscillator and the cathode-coupled multivibrator discharge circuits are then described, then synchronizing methods, waveform defects and their correction and, finally, deflection amplifiers. There are about 12 pages of general discussion on amplifiers and some 7 extra pages on line-scanning amplifiers in which attention is mainly directed to diode and triode damping methods. There are then 9 pages of typical commercial systems.

The book includes over 50 pages on projection television and a useful feature is a set of 48 oscillograms illustrating the waveforms and their amplitudes at various points, mainly in the time-base circuits, of a particular commercial receiver, the complete circuit of which is given. Although it is a projection receiver this has little effect on the circuitry and the waveforms are thus typical of most television sets. A point of particular interest is that the set includes a fly-wheel sync circuit and the waveforms in this are given.

In spite of the length of the book the treatment is not always as detailed as one could wish and there are quite a number of superficialities. Except in the last chapter, mathematics are excluded and the book concludes with a set of mathematical tables.

The book should prove extremely valuable to those for whom it is intended—that is, those who have to maintain and repair American television receivers. There is a good deal of material in it useful to those who handle British receivers. The book, however, can hardly be recommended to those without a fair knowledge already, for if they are beginners they will undoubtedly have difficulty in knowing which parts are relevant to British practice and which are not.

W. T. C.

Einführung in die Siebschaltungstheorie der elektrischen Nachrichtentechnik.

By R. Feldtkeller. Pp. 160 + viii, with 121 illustrations. S. Hirzel Verlag, Stuttgart. Price DM 12.

Dr. Feldtkeller is Director of the Telecommunications Institute of the Stuttgart Technische Hochschule. This is the third edition of this book on filter theory; the second edition in 1942 was a reprint of the first, but the present edition has been largely rewritten. In the introduction the author points out that this volume does not deal with high frequencies; he has another book "Theorie der Rundfunksiebschaltungen," the second edition of which appeared in 1944, which deals with high-frequency filters. The book starts with simple oscillatory circuits and gradually develops the various types of band-pass filters and filter chains. Their characteristics are calculated and plotted in a number of curves. Several sections are devoted to the m-derived filters of Zobel, and a final chapter deals with the effect of losses. The book concludes with an extensive bibliography but mainly of German publications. The treatment throughout is very thorough and the book is undoubtedly a valuable addition to the literature of the subject.

G. W. O. H.

Radio Servicing Equipment

By E. J. G. Lewis. Pp. 371 + xi, with 194 illustrations. Chapman & Hall, 37 Essex St., London, W.C.2. Price 25s.

Electrical Measurements and the Calculation of the Errors Involved

By D. Karo. Part I. Pp. 191, with 106 illustrations. Macdonald & Co. (Publishers) Ltd., 19 Ludgate Hill, London, E.C.4. Price 18s.

The Practical Electrical Reference Book

Pp. 384. Odhams Press Ltd., Long Acre, London, W.C.2. Price 9s. 6d.

ABSTRACTS and REFERENCES

Compiled by the Radio Research Board and published by arrangement with the Department of Scientific and Industrial Research

The abstracts are classified in accordance with the Universal Decimal Classification. They are arranged within broad subject sections in the order of the U.D.C. numbers, except that notices of book reviews are placed at the ends of the sections. U.D.C. numbers marked with a dagger (†) must be regarded as provisional. The abbreviations of the titles of journals are taken from the World List of Scientific Periodicals. Titles that do not appear in this List are abbreviated in a style conforming to it.

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ACOUSTICS AND AUDIO FREQUENCIES

534.22 : [546.212+546.212.02] 796

Velocity of Sound in Water and in Heavy Water as dependent upon Temperature.—P. P. Heusinger. (*Naturwissenschaften*, Sept. 1949, Vol. 36, No. 9, pp. 279–280.) Measurements were made at a frequency of 6.45 Mc/s in the temperature range 0°–100°C. The results are tabulated. For H_2O the velocity has a maximum value of 1 555.5 m/s at 73.4°C. For D_2O the maximum velocity (1 461.1 m/s) is at 75.1°C.

534.231 797

Elementary Treatment of the Fundamental Equations for the Radiation and Propagation of Sound.—F. A. Fischer. (*Frequenz*, Nov. 1949, Vol. 3, No. 11, pp. 320–328.) Discussion of (*a*) the two fundamental propagation equations, (*b*) spherical waves and their propagation through a sphere whose radius is small compared with the wavelength, (*c*) radiation from a piston membrane in an infinite rigid wall, (*d*) directed radiation, (*e*) radiation resistance.

534.231 798

Experimental Demonstration of a Theorem of Lord Rayleigh relative to the Production of Plane Waves.—J. Grunenwaldt. (*Ann. Télécommun.*, June 1949, Vol. 4, No. 6, pp. 203–209.)

534.6 : 621.395.623 : 534.771 799

Harmonic Distortion of Sound-Receivers and its Practical Consequences in Audiometry.—P. Chavasse, R. Caussé & R. Lehmann. (*Ann. Télécommun.*, May 1949, Vol. 4, No. 5, pp. 156–168.) The results of measurements of frequency response and harmonic distortion for high-quality receivers are presented in tables and curves as functions of frequency. The receivers included electrodynamic, piezoelectric and magnetic types. The response curves of the first two of these are regarded as satisfactory for audiometry purposes, though the piezoelectric receiver showed an antiresonance near 1 000 c/s. A telephone earpiece, however, showed a very pronounced resonance peak near 1 000 c/s, above which frequency the response decreased practically to zero at 5 000 c/s, making it quite unsuitable for any audiometry tests.

534.771 800

The Measurement of Hearing Loss.—J. T. Story. (*Marconi Instrumentation*, Nov./Dec. 1949, Vol. 2, No. 4, pp. 67–70.) General description of an audiometer, Type TF895, designed to the specification embodied in the Medical Research Council report on hearing aids (1245 of 1948). Ten tones in the range 125–8 000 c/s are provided, with simple intensity control which is calibrated relative to the normal threshold of hearing.

534.851+534.861.862+621.395.625.3 801

New Audio Trends.—J. M. (*Electronics*, Jan. 1950, Vol. 23, No. 1, pp. 68–71.) Three methods of synchronizing magnetic-tape records with cinema film are described. A new directional ribbon microphone, designed to eliminate the microphone-boom problem in television studios, has a pick-up distance of up to 12 feet and a frequency range of 50–15 000 c/s. A 78 r.p.m. disk-recording technique capable of reproduction up to 20 000 c/s is mentioned.

534.86 : 534.322.1 802

Influence of Reproducing System on Tonal-Range Preferences.—H. A. Chinn & P. Eisenberg. (*Proc. Instn Radio Engrs, Aust.*, Nov. 1949, Vol. 10, No. 11, pp. 297–306.) Reprint. See 2695 of 1948.

534.862.4 : 621.383.4 803

Lead-Sulfide Photoconductive Cells in Sound Reproducers.—Lee. (*See* 1027.)

534.862.4 : 621.395.625.3 804

The Reciprocal Relations between Magnetic Tape and Ring Head as affecting the Reproduction.—W. Guckenburg. (*Funk u. Ton*, Jan. 1950, Vol. 4, No. 1, pp. 24–33.) Discussion of the effect of the magnetic properties of the tape, and of the gap width used in the reproducing head, on the quality of the reproduction.

621.395.61 805

The Bantam Velocity Microphone.—L. J. Anderson & L. M. Wigington. (*Audio Engng, N.Y.*, Jan. 1950, Vol. 34, No. 1, pp. 12–14, 31.) Design of the KB-2C microphone is discussed; operational characteristics approximate to those of a conventional velocity microphone. Wind-screening, which is lower then normal due to proximity of screens to ribbon, is improved by addition of cotton or fibre-glass between the inner and outer screens, but with some resulting reduction in low-frequency response.

621.395.623.7 806

A Loudspeaker for the Range from 5 to 20 kc/s.—H. H. Smith & W. T. Selsted. (*Audio Engng, N.Y.*, Jan. 1950, Vol. 34, No. 1, pp. 16–18.) Theoretical design considerations and constructional details of a loudspeaker with a frequency response level to within ±2.5 db and an average efficiency of about 28%; it will handle a peak power of 5 W.

621.395.92 807

American Hearing Aids.—A. Dinsdale. (*Wireless World*, Jan. 1950, Vol. 56, No. 1, pp. 2–5.) Discussion of requirements and current practice in design and fitting. For comparable British information, see 2245 and 1246 of 1948.

AERIALS AND TRANSMISSION LINES

621.315.012.3 808

R.F. Transmission Line Nomographs.—P. H. Smith. (*Télévis. franç.*, Jan./Feb. 1950, No. 55, pp. 15–19.) French version of 1274 of 1949.

621.315.212 809

Reflection-Free Supporting Disks in Coaxial Cables.—H. Käden & G. Ellenberger. (*Arch. elekt. Übertragung*, Dec. 1949, Vol. 3, No. 9, pp. 313–322.) For non-reflection the diameter ratio for the disks must be made larger than that of the conductors. The outer and inner conductors must therefore be recessed where the supports fit. The necessary groove depth, which depends on the dielectric constant of the disk material, is calculated. To avoid field distortion the disks must be specially shaped. Exact mathematical determination of the optimum shape appears impracticable, but an approximate solution is obtained by a conformal transformation method. Experimental results confirm the theory.

621.317.336 : 621.396.67 810

Antennas and Open-Wire Lines: Part 3 — Image-Line Measurements.—P. Conley. (*J. appl. Phys.*, Nov. 1949, Vol. 20, No. 11, pp. 1022–1026.) An open-wire line is formed by a single wire and its image in a reflecting sheet. This line is inherently balanced and well-shielded. By loading at the centre and feeding at the ends, discontinuity at the load and the need for dielectric supports are avoided. Measurements on a symmetrical line give a simple method of determining the load impedance. A description is given of apparatus for laboratory investigation and of its application to the study of the coupling between a dipole and an open-wire line. Characteristics of the line compared favourably with those of a good coaxial measuring-line. Parts 1 and 2 : 281 and 399 of February.

621.317.336 811

Antenna Impedance Measurement by Reflection Method.—E. Istvánffy. (*Elect. Commun.*, Dec. 1949, Vol. 26, No. 4, pp. 285–291.) Reprint. See 2559 of 1949.

621.392.26† 812

Principal Waves in Electromagnetic Waveguides.—J. Ortusi & J. C. Simon. (*Ann. Radioélect.*, Jan. 1950, Vol. 5, No. 19, pp. 12–20.) The application of Maxwell's equations to guided waves is considered and the general solution is applied to the transverse e.m. wave as defined by a velocity of propagation equal to that of light in free space. The characteristic functions of such waves (power, voltage, current, characteristic impedance) are invariant in any conformal transformation of coordinates. This theorem is proved and then used to evaluate the iterative impedance of coaxial and multi-wire lines from the characteristic impedance of two strips of unbounded plane, which corresponds to the simplest T.E.M. wave.

621.392.26† 813

Waveguides (operating) beyond the Cut-Off Frequency: Application to Piston Attenuators.—A. Briot. (*Onde élect.*, Jan. 1950, Vol. 29, No. 274, pp. 57–63.) A detailed treatment is given of the characteristics of cylindrical piston attenuators for E_0 and for H_1 waves, and of rectangular attenuators for H_{01} waves. Matching of attennators to waveguides or coaxial lines for a given frequency band is also discussed. See also 551 of March.

621.392.26† : 621.396.67 814

Slotted Waveguides and their Application as Aerials.—M. Bouix. (*Ann. Télécommun.*, March 1949, Vol. 4, No. 3, pp. 75–86.) A transmission line with 'small loads' distributed along it is first discussed. These loads are in the form of quadripoles whose shunt admittance and series impedance are small compared with the characteristic impedance of the line. Representation of such loaded lines by means of circle diagrams leads to the concept of an equilibrium cycle. A study is made of resonant slots in the wall of a waveguide, these slots constituting the 'small loads'. The disposition of slots for a desired radiation distribution is considered, in particular that giving (*a*) uniform distribution and (*b*) a 1 : 4 : 1 'gable' distribution. The relation between squint angle and the spacing of the slots is determined for different waveguides of American standard dimensions. A method of designing slot systems is outlined and the case is considered of a system of slots with $\lambda_g/2$ spacing, the waveguide being terminated by a short-circuiting plunger at a distance of $\lambda_g/4$ from the end slot. Applications of slotted aerials are briefly reviewed.

621.392.43.012.2 815

A Graphic Procedure for Calculations involving Transmitting-Line Systems.—de Onis. (*See* 837.)

621.396.67 816

Application of a Variational Principle to Biconical Antennas.—C. T. Tai. (*J. appl. Phys.*, Nov. 1949, Vol. 20, No. 11, pp. 1076–1084.) A theoretical examination is made of the impedance of a biconical aerial, using a method devised by Schwinger. An integral equation for the aperture field is obtained by fulfilling boundary conditions on a sphere. Hence an expression is derived for the effective terminating admittance. The solution of zero order is shown to be identical with that obtained by Smith by neglecting higher modes than the principal mode in the interior of the boundary sphere. For cones of small angle, the result agrees with exact solutions obtained by other methods. For cones of any angle, a first-order solution is obtained, and realized numerically for certain angles. See also 1588 of 1949.

621.396.67 817

Omnidirectional Wide-Band Aerials for Decametre and Metre Waves.—O. Zinke. (*Radio franç.*, Dec. 1949, No. 12, pp. 7–12.) The construction and impedance characteristics of different types of such aerials are considered. For a frequency coverage $f_{min} : f_{max}$ of 1 : 2 or 1 : 3, the diameter/length ratio should be large (about 0.5). The use of plates, tubes, wire mesh, and multi-wire aerials to obtain such a ratio is illustrated.

621.396.67 : 621.316.761.2 818

Wide-Band Aerials and Resonant Circuits with Simple and Double Compensation.—O. Zinke. (*Radio franç.*, Dec. 1949, No. 12, pp. 13–17.) The compensating reactance ratios for series and parallel circuits are determined. A s.w.r. of 2.6 can be reduced to 1.4 by the use of one equalizing circuit, and to 1.25 by using two

circuits. Their application to a fan-type aerial of length $< \lambda/4$ is considered, and shows that the s.w.r. may be kept constant for variations of frequency from -16% to $+14\%$ of the resonance frequency. Near *antiresonance a much wider frequency band* is obtained, with a similar s.w.r. The equivalence of these compensating circuits to T and π filters is noted.

621.396.67 : 621.397.62 819

Built-In Antennas for Television Receivers.—K. Schlesinger. (*Electronics*, Jan. 1950, Vol. 23, No. 1, pp. 72–77.) Electrical designs for three types of built-in aerials are given: a short dipole, a bi-resonant loop and a double loop. The characteristics of the aerials are described and their performance is compared with that of a $\lambda/2$ dipole. Measured values of signal attenuation at various locations in a 3-storey brick/steel building are given for three frequencies.

621.396.671 820

Gain of Aerial Systems.—D. A. Bell. (*Wireless Engr*, Jan. 1950, Vol. 27, No. 316, p. 32.) Comment on 287 of February (Brown).

621.396.677 821

Aluminium aids Television.—(*Metal Ind., Lond.*, 27th Jan. 1950, Vol. 76, No. 4, pp. 71–73.) Details are given of the construction of the 14-ft paraboloid grid-type reflectors for the London/Birmingham link. These are constructed from Al-alloy tubes and castings, the tubes forming the actual reflectors being $\frac{3}{8}$ in. in diameter and spaced 3 in. between centres. Insulated heater cables are fitted inside the reflector tubes to prevent icing, the heater system being divided into four sections; maximum power dissipated in either the two inner or the two outer sections is 6 kW.

621.396.67 822

Aerials for Metre and Decimetre Wavelengths. [Book Review]—R. A. Smith. Publishers: Cambridge University Press, London, 218 pp., 18s. (*Wireless Engr*, Jan. 1950, Vol. 27, No. 316, pp. 30–31.) "The aerial systems described are mainly those which have been developed for the wavelength range of 12 m to 1 m, and only one short chapter is devoted specifically to decimetre-wave aerials. . . . a most lucid and valuable account of the subject. The book . . . may be unreservedly recommended."

CIRCUITS AND CIRCUIT ELEMENTS

621.3.015.3 : [517.63 + 517.432.1 823

The Laplace Transformation and the Study of Transient Phenomena.—Colombo. (*See* 914.)

621.3.016.35 824

On the Stability Criterion of H. Nyquist.—F. Kirschstein. (*Arch. elekt. Übertragung*, Sept. 1949, Vol. 3, No. 6, pp. 195–198.) A discussion of Nyquist's rule that an automatic regulating device is stable against self-excitation if its "C-curve" does not embrace the critical point (1, 0) in the complex plane. It is proved by an example that this rule does not always hold, and a practical criterion of stability is given.

621.3.016.35 : 621.3.012.2 825

Practical Stability Testing by means of Circle Diagrams.—F. Strecker. (*Elektrotechnik, Berlin*, Dec. 1949, Vol. 3, No. 12, pp. 379–388.)

621.314.2.012.3 826

Abacs for Output-Transformer Design.—P. L. Courier. (*T.S.F. pour Tous*, Jan. 1950, Vol. 26, No. 225, pp. 13–17.) The first abac gives the transformation and load distribution for optimum impedance matching; the second presents sets of curves from which l.f. and h.f. attenuation and the circuit constants can be determined. Practical numerical examples are given.

621.314.26 827

On the Frequency Conversion of Technical Alternating Currents to High Frequency by means of Valves.—E. Prokott. (*Fernmeldetech. Z.*, Oct. 1949, Vol. 2, No. 10, pp. 301–308.) A push-pull transmitter-amplifier is described which is fed with a.c. and gives a h.f. output modulated by a frequency which is twice the supply frequency. With two or more amplifier stages, good efficiency can be obtained. If high-emission valves are used, considerable power can be handled. The transmitter can be keyed and ph.m. or f.m. can be applied. If tone-frequency energy is used for supplying a push-pull stage, sidebands can be produced.

621.316.761.2 : 621.396.67 828

Wide-Band Aerials and Resonant Circuits with Simple and Double Compensation.—Zinke. (*See* 818.)

621.318.4 829

The Frequency Dependence of the Distortion for Coils with Laminated-Iron Cores.—H. Kämmerer. (*Arch. elekt. Übertragung*, Oct. 1949, Vol. 3, No. 7, pp. 249–256.) Starting from the Rayleigh hysteresis formula and eddy current theory, the frequency characteristic of the distortion factor is determined graphically. For very small field strengths, the calculated values agree well with measurements of the distortion for permenorm under no-load conditions. For larger field strengths, the measured distortion is less than the calculated value.

621.318.572 830

A Stepping Scale-of-Ten Counting Unit.—I. A. D. Lewis & J. F. Raffle. (*J. sci. Instrum.*, Jan. 1950, Vol. 27, No. 1, pp. 7–10.) The principle, operation and performance of the circuit are discussed in some detail with the aid of a circuit diagram and waveforms. Developed along the lines of the Miller integrator, the circuit comprises two pentodes and five diodes. It is primarily designed for connection to the output of a commercial scale-of-100 unit. Counting is achieved by the collection of discrete quantities of charge on a capacitor. The mechanical counter sets an upper limit of 9 counts per sec at the output.

621.318.572 831

Electronic Amplitude Selector with a Scaling-Down Circuit.—T. Kahan & A. Kwartiroff. (*J. Phys. Radium*, Dec. 1946, Vol. 7, No. 12, pp. 357–359.) Description with complete circuit diagram of apparatus which will select and record pulses irrespective of pulse shape, phase or duration. The associated *n*-scale circuit controls the number recorded according to the scale set.

621.318.572 : 621.384.5 832

Neon Diode Ring Counter.—J. C. Manley & E. F. Buckley. (*Electronics*, Jan. 1950, Vol. 23, No. 1, pp. 84–87.) An inexpensive 3-decade instrument with a counting speed up to 500 impulses per sec. Neon glow-discharge tubes and Ge diodes are used. Circuit details and the mode of operation are described. The counter can be reset instantaneously and the action can be reversed for subtraction. An arrangement is included which produces a special signal when any selected number up to 999 is reached.

621.392 833

Network Theorem.—V. Belevitch. (*Wireless Engr*, Jan. 1950, Vol. 27, No. 316, p. 33.) The theorem discussed by Wigan (305 of February) has been proved by Cauer (392 of 1942) and applied to various problems in filter design.

834

621.392

Electrical Analogs of Linear Systems.—J. P. Corbett. (*Elect. Engng, N.Y.*, Dec. 1949, Vol. 68, No. 12, p. 1075.) Summary only. Discussion of transformations which can be used to change a linear system into one capable of being more easily represented electrically.

835

621.392 : 621.396.813

Some Aspects of the Theory of Bode.—H. Familier. (*Ann. Radioélect.*, Jan. 1950, Vol. 5, No. 19, pp. 36–53.) The general principles of the theory (1381 of 1948) are summarized and three basic relations between the real and imaginary components of the principal circuit function $\theta = A + jB$ are derived. These expressions are developed further to determine the attenuation curve corresponding to a given phase curve in the study of distortion in a selective circuit. The theoretical relations are then applied to negative-feedback amplifiers to determine their characteristics (gain, impedances, etc.) and stability. As an example, a wide-band video amplifier is studied and the optimum total feedback is calculated from given specifications. Two appendices deal respectively with (*a*) theorems relating to the physical realizability of circuits, (*b*) integration in the complex plane, with application of Cauchy's theorem.

836

621.392.43

The Exponential Line as a Transformer.—O. Zinke. (*Radio franç.*, Jan. 1950, No. 1, pp. 9–13.) See 672 of 1948.

837

621.392.43.012.2

A Graphic Procedure for Calculations involving Transmitting-Line Systems.—A. de Onis. (*Commun. News*, Oct. 1949, Vol. 10, No. 3, pp. 87–97.) A detailed account of the practical use of circle diagrams in calculations involving such devices as matching stubs or transmission-line transformers. Symmetrical branching of the line is also considered. The line is assumed to be lossless and to have a purely resistive characteristic impedance. These conditions are usually satisfied approximately by r.f. feeders, for which the methods described are particularly intended.

838

621.392.5

On Quadripoles having their Iterative Impedances Proportional to Those of a Given Quadripole.—M. Parodi. (*J. Phys. Radium*, May 1947, Vol. 8, No. 5, pp. 140–142.)

839

621.392.5

Analytical Conditions for Damping in an Electrical Network of *n* Independent Meshes. (Note on Hurwitz Polynomials in the Form of Determinants).—M. Parodi & F. Raymond. (*Ann. Télécommun.*, June 1949, Vol. 4, No. 6, pp. 231–232.) The necessary and sufficient conditions are determined for a symmetrical determinant, with real elements, of order n, $|a_{ik}z^2 + b_{ik}z + c_{ik}|$, to be a Hurwitz polynomial in z. A classical result in the theory of electrical networks is thus established. This result is generalized.

840

621.392.5.018.1 : 621.396.615

The Design of RC Oscillator Phase-Shifting Networks.—W. R. Hinton. (*Electronic Engng*, Jan. 1950, Vol. 22, No. 263, pp. 13–17.) A method of applying matrix algebra to determine the coefficients used in the equations giving the operating frequency and the gain required to maintain oscillation in RC oscillators. Coefficients for the design equations for a number of different networks are given.

841

621.392.52

Design of Absorption Traps.—J. Avins. (*Electronics*, Jan. 1950, Vol. 23, No. 1, pp. 105–108.) Universal response curves show the ratio of the response of a tuned circuit, to which a trap circuit is inductively coupled, to the response without the trap, for typical values of attenuation and frequency separation. An abac for determination of the coupling factor is given. The response curves indicate that optimum performance is obtained when an absorption trap is coupled to a circuit which is relatively close in frequency.

842

621.392.52

Calculation of Band-Pass Filters using Piezoelectric Crystals in Lattice Structures.—A. Fromageot & M. A. Lalande. (*Elect. Commun.*, Dec. 1949, Vol. 26, No. 4, pp. 305–318.) Theory is developed and applied to the practical design of wide- and narrow-band filters. The attenuation curves are given of a filter for separating the channels in a s.s.b. R/T system, and also for the pilot-selecting filter for the receiver used in that system.

843

621.392.52

A Three-Stage High-Frequency Filter with Variable Bandwidth, and its Practical Realization.—W. Pfost. (*Arch. elekt. Übertragung*, Sept. & Oct. 1949, Vol. 3, Nos. 6 & 7, pp. 199–202 & 257–264.) The construction of the filter is described and performance figures are discussed. These indicate that the selective gain of such a filter is nearly independent of the effective bandwidth. With given maximum Q values for coils and circuit, maximum gain is obtained for $N = 5$, where N is the ratio of the sum of the loss angles of all circuits to that of the best circuit. It is advantageous to deviate slightly from the shape of the exact Tchebycheff attenuation curve within the pass band. Filters with N from 4 to 4.5 can then be designed with the same bandwidth and a sharper cut-off, without appreciable reduction in gain. Experiments confirm these conclusions.

844

621.392.52 : 621.3.094.1'.2 : 621.396.813

Modulation Distortion by Band-Pass Filters.—U. Finkbein. (*Funk u. Ton*, Jan. 1950, Vol. 4, No. 1, pp. 11–15.) Attenuation- and phase-distortion of the modulation frequency in a h.f. filter are calculated, for the steady state, from the amplitude and phase characteristics of the filter. With a symmetrical filter curve and accurate tuning, distortion is at a minimum. The use of a low modulation depth and quadratic rectification is also advantageous.

845

621.392.52 : 621.396.823

The Design of L-Type Inductance-Capacitance High-Frequency Filters for Suppressing Industrial Interference.—S. A. Lyutov. (*Radiotekhnika, Moscow*, Sept./Oct. 1949, Vol. 4, No. 5, pp. 28–40. In Russian.)

846

621.392.53

The Design of Reactive Equalizers.—A. P. Brogle, Jr. (*Bell Syst. tech. J.*, Oct. 1949, Vol. 28, No. 4, pp. 716–750.) "This paper describes a systematic method of approximating with a finite number of network elements a transfer characteristic which is a prescribed function of frequency, rather than a constant, over the useful frequency band. Although applied here only to input and output coupling networks as reactive equalizers and where loss equalization to an extremely high degree of precision over a wide frequency band is desired, the mathematical expressions which form the basis for the design are applicable to any 4-terminal network whose transfer characteristic is specified in a similar manner over the real frequency range.

"The selection of the appropriate form of the transfer function for equalization purposes is the fundamental consideration. A squared Tchebycheff polynomial is found to be particularly suitable to produce a desired cut-off characteristic without impairing the precision of equalization in the useful band.

"A method of polynomial approximation based on the transformation $\omega = \tan \phi/2$ is used to obtain the coefficients of the in-band approximating function. Predistorting the transfer specification and minimizing

the mean-square error, the coefficients become the Fourier cosine coefficients for an infinite frequency range, and are the solutions of a linear set for a finite range, $0 \leq \phi \leq \pi/2$."

847

621.396.611.1 : 517.93

Harmonic and Subharmonic Response for the Duffing Equation.—M. E. Levenson. (*J. appl. Phys.*, Nov. 1949, Vol. 20, No. 11, pp. 1045–1051.) The Duffing equation $\ddot{x} + \alpha x + \beta x^3 = F \cos \omega t$, in which $\alpha > 0$, possesses periodic solutions with frequency ω/n for all integral values of n, provided β is sufficiently small. Two types of harmonic solutions and two types of subharmonic solutions exist if β is small enough. The perturbation method is used to find the approximate response curves for each of the four types of solution. The response curves obtained in the nonlinear case ($\beta \neq 0$) and in the linear case ($\beta = 0$) are compared and some properties of the solutions in the nonlinear case are discussed. For one particular type of subharmonic solution a comparison is made between the perturbation method and the iteration method used by Rauscher, in which F is assumed small instead of β. The main result obtained is that the two methods yield similar results for larger values of β than might be expected. See also 582 of March (Elias & Duinker).

848

621.396.615 : 621.392.52

Oscillators of High Frequency-Constancy.—W. Herzog. (*Arch. elekt. Übertragung*, Sept. 1949, Vol. 3, No. 6, pp. 203–207.) The Q value of an oscillator is defined in terms of the product of the feedback and amplification factors. Oscillators developed from filter circuits can be much improved by increasing the amplification and also by increasing the steepness of the sides of the filter attenuation curve. Improved forms of oscillators of the bridge and Heegner types are discussed.

849

621.396.615.029.63

On a Particular Characteristic of Disk-Seal Valves 2C40 and 2C43.—Liot. (*See* 1033.)

850

621.396.615.14

Coaxial Oscillator Circuit for Decimetre Waves.—L. Liot. (*Radio franç.*, Jan. 1950, No. 1, pp. 14–17.) A scale diagram and description of a 600-Mc/s oscillator giving a useful output of 18 W. The oscillator comprises a grounded-grid triode Type T30H (783 of March) with a $\lambda/2$ coaxial-line anode circuit and a $3\lambda/4$ coaxial-line cathode circuit. Maximum operating frequency is 714 Mc/s. Practical rules for simplifying design calculations for this type of oscillator will be given later.

851

621.396.615.17

A Flexible High-Voltage Square-Wave Generator.—L. C. Hedrick. (*Rev. sci. Instrum.*, Nov. 1949, Vol. 20, No. 11, pp. 781–783.) Description of a generator suitable for modulating the absorption cell of a microwave spectrograph. Peak-to-peak voltages up to 1 800 V have been produced at 100 kc/s, feeding into a capacitive load of 800 pF. Voltage adjustment is effected by variation of a single component. The instrument is kept effectively at ground potential automatically.

852

621.396.615.17

Transitron Relaxation Oscillators and the Transitron Integrator.—J. Moline. (*Radio franç.*, Dec. 1949, No. 12, pp. 20–23.) The principles of operation of the transitron relaxation circuit and of the Miller integrator are described. The combination of the two systems in one circuit is illustrated. This circuit generates a pure sawtooth waveform with good stability. Component values for such a circuit using an EF50 valve are noted.

853

621.396.615.17

Variable-Pulse-Length Generator.—J. C. May. (*Electronics*, Jan. 1950, Vol. 23, No. 1, pp. 109–111.) A cathode-coupled clipper circuit with regeneration provides pulses of width linearly variable from 0.5 μs to 24 μs and with peak voltages of 4.5–6.5 V.

854

621.396.619.16

Development in Pulse-Modulation Circuits.—F. Butler. (*Wireless Engr.*, Jan. 1950, Vol. 27, No. 316, pp. 12–16.) The pulses are generated across the common anode load of two valves fed in push-pull from a low-power primary oscillator whose frequency equals the desired recurrence frequency. Reactance modulator valves are used to vary the phase of the grid-cathode voltage of one valve with respect to that of the other, giving width-modulation of the pulses. The system is suitable for speech transmission and can be adapted to give several channels on a time-allocation multiplex basis.

855

621.396.621.5

Methods of Calculating the Sensitivity of Mixer Stages for Decimetre and Centimetre Waves.—H. F. Mataré. (*Arch. elekt. Übertragung*, Oct. 1949, Vol. 3, No. 7, pp. 241–248.) A knowledge of diode theory is assumed; previous work on the subject is reviewed and methods of calculating sensitivity are discussed. The application of these methods in the case of diode mixing is considered. A mathematical analysis shows that by introducing mean values into the expression for the noise function of the network, calculation of sensitivity and receiver circuit design are simplified. A numerical treatment of the mixer-stage equation, together with a comparison of calculated and experimental results, will be given in a later paper.

856

621.396.645

Design of an Amplifier with Circuits Tuned to Different Frequencies and such that the Phase/Frequency Function is Linear in a Given Range of Frequencies.—A. Fromageot & P. Belgodère. (*Ann. Télécommun.*, May 1949, Vol. 4, No. 5, pp. 169–176.) Expressions for the modulus and phase of the transfer constant of an n-stage stagger-tuned amplifier are derived which involve the Newtonian potential function and its conjugate. Certain properties of these functions are reviewed, in particular their behaviour with regard to conformal transformation. The continuous distribution of charge which corresponds to a linear variation of the phase function with frequency in a limited frequency band is considered, and an estimate is made of the order of approximation resulting from the adoption of a discontinuous distribution of point charges. The method of calculation enables the tuning frequencies and the Q values of the different circuits to be determined for an amplifier of given gain in a given frequency band.

857

621.396.645

High-Quality Amplifier.—D. T. N. Williamson. (*Wireless World*, Jan. 1950, Vol. 56, No. 1, p. 24.) Replies to readers' queries on 3101 of 1949 and 333 of February.

858

621.396.645

Reducing the Effect of Capacitance in Screened Cable.—G. H. Rayner. (*Electronic Engng.*, Jan. 1950, Vol. 22, No. 263, p. 33.) The input capacitance at a.f. of a cathode follower with customary connections is about 10 pF. To reduce this the valve should be mounted within a screen that can be connected to the cathode and the screen-grid should be coupled to the cathode and not by-passed to earth. The input capacitance can thus be reduced by a factor of about 10.

621.396.645 : 621.392.52 **859**

RC-Amplifier Filters.—C. H. Miller. (*Wireless Engr.*, Jan. 1950, Vol. 27, No. 316, pp. 26–29.) "Simple expressions are derived for the approximate cut-off frequencies of amplifiers which use resistance-capacitance feedback networks to produce high-pass, low-pass, or band-pass frequency characteristics similar to those of conventional filters. The design of such amplifiers is discussed, and illustrations are given of the results obtained with simple amplifiers of this type."

621.396.645 : 621.392.52 **860**

On the Problem of Bandwidth Control.—J. Harmans. (*Funk u. Ton*, 1949, Vol. 3, Nos. 11/12, pp. 570–574.) The conditions necessary for an optimum frequency/amplification response curve (see 1889 of 1948) are realized by the use of two 2-stage filters with variable coupling. Sensitivity, selectivity and bandwidth are considered for two cases: (*a*) with one filter variable, the other fixed; (*b*) with both variable. With an overall bandwidth of 21.7 kc/s, a deviation of less than ± 5% from the ideal response curve is obtainable.

621.396.645 : 621.396.611.3 **861**

On the Connection between Amplification and Bandwidth in a Multi-Stage Amplifier with Circuits of the Same Resonance Frequency.—W. Kleen. (*Funk u. Ton*, 1949, Vol. 3, Nos. 11/12, pp. 584–591.) The circuit considered uses a single type of valve with given characteristics. The valves are connected in cascade and are coupled by simple rejection circuits. A formula is derived which gives the maximum attainable amplification for a given bandwidth, and the corresponding values are determined for number of stages, amplification per stage, load resistance and bandwidth for a single stage. A numerical example is calculated. Similarly, the maximum bandwidth for a given overall amplification is determined.

621.396.645.37 : 621.3.016.35 **862**

Investigation of the Conditional Stability of Feedback Amplifiers.—K. H. R. Weber. (*Frequenz*, Sept. 1949, Vol. 3, No. 9, pp. 253–259.) A theoretical and experimental investigation of the dependence of the stability on the parameter vK, where v is the voltage gain of the amplifier without feedback and K is the gain in the feedback path. For the amplifier used, conditional stability existed for values of vK between 4 000 and 100 000. The conditions for instability to exist and hence for self-oscillation to occur when switching on or off are summarized. The smaller the distance of the vK characteristic from the point (1, 0), the lower is the maximum permissible input voltage for self-oscillation not to occur. With the experimental amplifier, self-excitation commenced at an input voltage of 700 mV for $v = 100\,000$, $K = 1$, but at 180 mV for $v = 10\,000$, $K = 0.5$.

621.396.645.371 **863**

Negative-Feedback Amplifiers.—É. Chamagne & G. Guyot. (*Ann. Télécommun.*, April 1949, Vol. 4, No. 4, pp. 119–122.) Formulae are derived for the amplification factor and output impedance of a feedback amplifier and also for the transient response. Results are given for an amplifier designed to transmit a band of width 3 Mc/s.

621.396.662 **864**

A High-Power Attenuator for Microwaves.—D. Alpert. (*Rev. sci. Instrum.*, Nov. 1949, Vol. 20, No. 11, pp. 779–781.) The attenuator is for use with waveguides working at λ 10 cm. The attenuating material is a mixture of ethylene glycol and water which serves also to remove the absorbed energy. The variations in attenuation and power s.w.r. with mixture concentration are shown. Power up to 100 W can be absorbed. Attenuation is variable from 1.5 db to 40 db and power s.w.r. is less than 1.1 at all settings.

621.392 **865**

Networks, Lines and Fields. [Book Review]—J. D. Ryder. Publishers: Prentice-Hall, New York, 1949, 462 pp., $7.35. (*Electronics*, Jan. 1950, Vol. 23, No. 1, pp. 208, 210.) "A rather complete treatment on the undergraduate level of elementary network theory, filters, transmission lines at low and high frequencies, Maxwell equations, reflection of waves, and waveguides." "... will be useful not only to undergraduates but also to an average engineer for direct consultation."

621.396.662.3 + 621.396.645 + 621.396.62 **866**

Applications of the Electronic Valve in Receiving Sets and Amplifiers. [Book Review]—Dammers, Haantjes, Otte & van Suchtelen. (*See* 991.)

GENERAL PHYSICS

53.081 + 537.713 **867**

The Rationalized Giorgi System and its Consequences.—P. Cornelius & H. C. Hamaker. (*Philips Res. Rep.*, April 1949, Vol. 4, No. 2, pp. 123–142.)

530.12 **868**

Equations of Motion and Tensor of Matter for the System of Finite Mass in the General Theory of Relativity.—N. M. Petrova. (*Zh. eksp. teor. Fiz.*, Nov. 1949, Vol. 19, No. 11, pp. 989–990. In Russian.)

534 **869**

On Certain Cases of the General Problem of the Theory of Steady-State Oscillations.—D. I. Sherman. (*C. R. Acad. Sci. U.R.S.S.*, 21st Aug. 1947, Vol. 56, No. 6, pp. 567–570. In Russian.) An investigation of solutions of the two-dimensional wave equation which satisfy mixed boundary conditions containing derivatives of the wave function up to any finite order. The problem depends on the solution of an integral equation.

535.43 **870**

The Scattering Cross Section of Spheres for Electromagnetic Waves.—L. Brillouin. (*J. appl. Phys.*, Nov. 1949, Vol. 20, No. 11, pp. 1110–1125.) For the case of large spheres, where geometrical optics should apply, the rigorous classical theory yields a scattering cross-section equal to twice the actual cross-section of the sphere. The same holds good for large cylinders. An explanation of this anomalous result is given which shows the role played by the shadow and by the diffraction fringes surrounding the shadow. A reasonable system of approximations yields the well-known Babinet's principle. The physical interpretation is so general that it must apply to similar problems in acoustics or wave mechanics. Experiments designed by Sinclair and LaMer gave an accurate check of the theory.

537.291 : 537.525.92 **871**

Electron Flow in Curved Paths under Space-Charge Conditions.—B. Meltzer. (*Proc. phys. Soc.*, 1st Dec. 1949, Vol. 62, No. 360B, pp. 813–817.) A theoretical treatment of a particular example of 3-dimensional flow for which the electron trajectories all lie on hyperboloids and the equipotential surfaces are ellipsoidal. Some general theorems are deduced from the results of an earlier paper (3120 of 1949). Distinction is made between "normal" flow, in which all electrons on a given equipotential have the same speed, and "abnormal" flow in which the speeds are not the same. It is shown that in normal flow the velocity vector obeys the law $\boldsymbol{v} \times \operatorname{curl} \boldsymbol{v} = 0$.

537.311.62 **872**

Anomalous Skin Impedance as a Function of the Field Strength.—K. F. Niessen. (*Philips Res. Rep.*, April 1949, Vol. 4, No. 2, pp. 143–153.) A study of the skin effect in a metal of high conductivity subjected to a strong h.f. field. See also 3403 of 1949.

537.525 873

Breakdown in Cold-Cathode Tubes at Low Pressure.—F. G. Heymann. (*Proc. phys. Soc.*, 1st Jan. 1950, Vol. 63, No. 361B, pp. 25–41.)

537.525.92 : 621.385.029.63 : 64 874

Increasing Space-Charge Waves.—J. R. Pierce. (*J. appl. Phys.*, Nov. 1949, Vol. 20, No. 11, pp. 1060–1066.) The equation for the gain per unit length of increasing waves in the presence of two streams of charged particles has been solved numerically for a wide range of values of plasma frequencies and stream velocities. The solutions are presented as curves of a gain parameter versus the ratio of the frequency to the plasma frequency of the faster stream, for various ratios of the stream velocities. Each graph corresponds to a different ratio of plasma frequencies. These curves apply to such effects as ion noise and double-stream r.f. amplification and it is shown that in certain circumstances the latter may occur in valves intended to have only a single electron stream. The case of streams moving in opposite directions is also considered. See also 3406 of 1949 (Bailey) and back references, 2486 of 1949 (Pierce & Hebenstreit).

537.562 875

Measurement of the Complex Conductivity of an Ionized Gas at Microwave Frequencies.—F. P. Adler. (*J. appl. Phys.*, Nov. 1949, Vol. 20, No. 11, pp. 1125–1129.) The positive column of a glow discharge is placed along the axis of a cylindrical cavity excited in the TM_{010} mode. The transmission of 3-cm waves through the cavity and the shift in resonance frequency are observed as a function of discharge current. It is shown that from these measurements values of the complex conductivity, $\sigma_r + i\sigma_i$, of the electron gas can be calculated. Curves of the measured conductivity components as functions of pressure and current are given. Using a theoretical formula for the conductivity (Margenau: 3246 of 1946) values of electron density can in turn be calculated from both σ_r and σ_i. Langmuir probe studies are carried out to check the results obtained, and adequate agreement is found.

538.56 : 537.71 876

Concerning 'The Impedance of Free Space'.—A. Foch. (*Onde Elect.*, Jan. 1950, Vol. 29, No. 274, pp. 5–7.) Comment on the anomalies of the two systems of units, classical and rationalized, with an explanation of the apparent discrepancy between the values of free-space impedance for the two systems. See also 607 of March (Budeanu) and 877 below.

538.56 : 537.71 877

On the Impedance of Free Waves.—E. Brylinski. (*Bull. Soc. franç. Elect.*, Jan. 1950, Vol. 10, No. 100, pp. 37–40.) The ratio Z_0 of the e.s. field $\mathcal{E}$ to the e.m. field H of a free-space e.m. wave has the dimensions of a resistance but is not a true impedance. It is suggested that it should be called the pseudo-impedance of free space and measured in pseudohms. In the classic (non-rationalized) Giorgi system Z_0 would have the value 30 pseudohms. In the rationalized Giorgi system $Z'_0 = 4\pi Z_0 = 377$ pseudohms and Z'_0 can be distinguished by calling it the pseudo-impedance of free waves. It should be noted that Z_0, like the velocity c to which it is closely related, can be either a scalar or a vector quantity.

539.11 878

Some Electronic Properties of a One-Dimensional Crystal Model.—D. S. Saxon & R. A. Hutner. (*Philips Res. Rep.*, April 1949, Vol. 4, No. 2, pp. 81–122.) "The Kronig-Penney model of a one-dimensional crystal is studied further with the aid of the formalism of the Dirac δ-function. Both a Green's-function method and a scattering-matrix method are used to derive the energy levels and wave functions for the monatomic and the diatomic lattices. The scattering-matrix method is used to study the problems of a single impurity, both substitutionally and interstitially located, and the coupling between impurities. In the last section the general problem of a solid solution is discussed."

548.0 : 539.11 : 537.529 879

On the Theory of Electron Multiplication in Crystals.—F. Seitz. (*Phys. Rev.*, 1st Nov. 1949, Vol. 76, No. 9, pp. 1376–1393.) The non-polar coupling between electrons and lattice has an important effect on the retardation of electrons. This retardation is a maximum when the energy of the electron corresponds to the energy level near the boundary of the Brillouin zone. The retardation arising from purely polar interaction in polar crystals is a maximum when the energy is near that of the polar modes. Statistical fluctuations in the velocity of the electrons largely determine electron multiplication, and the value of the field producing breakdown in a standard specimen is no more than a fifth of that obtained from the criteria of von Hippel and Fröhlich.

621.3.011.4 : 513.466 880

Potential Problems and Capacitance for a Conductor Bounded by Two Intersecting Spheres.—C. Snow. (*Bur. Stand. J. Res.*, Oct. 1949, Vol. 43, No. 4, pp. 377–407.) Formal expressions, as series and integrals, are derived for the external potential, which takes on assigned values on the two spherical boundary surfaces, either when the body is alone or when it is in any given axially symmetric e.s. field. This is effected by representing a given function $f(x)$, for $1 < x < \infty$, as a complex integral whose variable is the parameter μ of a Legendre function $P_{\mu-\frac{1}{2}}(x)$ or $Q_{\mu-\frac{1}{2}}(x)$. The capacitance of the conductor is found as a series whose terms involve psi-functions. In case the two spheres intersect at an angle ω which is a rational function of 2π, the capacitance is given in finite terms involving complete elliptic integrals. The field or e.s. potential is given in finite terms when $\omega = n\pi/m$, where m is any positive integer, and $n = 1$, 2, 3 or 4. The cases $n = 3$, $n = 4$ involve elliptic functions. These would permit the exact computation in finite terms of the penetration of an external applied field into a cavity with any angular aperture.

53 881

The Physical Society Reports on Progress in Physics. Vol. 12 (1948–49). (Book Review)—Publishers: The Physical Society, London, 42s. (*Phil. Mag.*, Nov. 1949, Vol. 40, No. 310, p. 1178.) "The Physical Society has been remarkably successful in obtaining excellent articles on subjects of current interest from authors of eminence." A list of the articles, with authors, is given.

534.1 + 538.56 882

Einführung in die Lehre von den Schwingungen und Wellen. (Introduction to the Theory of Vibrations and Wave-Motion) (Book Review)—K. W. Wagner. Publishers: Dieterichsche Verlagsbuchhandlung, Wiesbaden, 1947, 640 pp., 34 DM. (*Arch. elekt. Übertragung*, Sept. 1949, Vol. 3, No. 6, pp. 230–231.) A second edition of the full course of lectures given by the author in the Technische Hochschule, Berlin. It is a book for the rising engineer and physicist, with a wealth of material for the teacher and professional man. Aspects dealt with are: forms of oscillation; dynamics of vibration; oscillations and waves in continuous systems; e.m. waves in the ionosphere; vibration systems; nonlinear systems. Examples of both mechanical and electrical oscillations are considered and the two types related. The book is considered by the reviewer to fill a former gap in the literature on the subject.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.72 : 621.396.822 883

Solar Radio-Frequency Radiation of Thermal Origin.—J. L. Pawsey & D. E. Yabsley. (*Aust. J. sci. Res., Ser. A*, June 1949, Vol. 2, No. 2, pp. 198–213.) Numerous measurements of 'noise flux density' in the wavelength range from 1 cm to 4 m are examined. In addition to the occasional bursts, a relatively constant component, which has the characteristics of thermal radiation, is always found; its intensity increases uniformly with frequency from about 10^4 °K at 1.25 cm to about 10^8 °K at 1.5 m, in substantial agreement with Martyn's theory (2783 of 1948).

523.72 : 621.396.822 884

Bursts of Solar Radiation at Metre Wavelengths.—R. Payne-Scott. (*Aust. J. sci. Res., Ser. A*, June 1949, Vol. 2, No. 2, pp. 214–227.) The variable high-intensity radiation on 85, 65, 60, and 19 Mc/s was studied during part of 1948. Two types were observed: (*a*) circularly polarized long-duration radiation; (*b*) short unpolarized bursts. The bursts on 85 Mc/s arrive 0.7 sec before, and on 19 Mc/s 9 sec after the corresponding bursts on 60 Mc/s; they decay exponentially with a time constant of about 1.7 sec; double-humped bursts are common. Their characteristics agree with the hypothesis that they originate in the high corona, that the time constant depends on the collision frequency, and that the second hump is due to a reflection from the inner corona. The larger bursts are usually associated with solar flares.

523.72 : 621.396.822 885

The Noise-Like Character of Solar Radiation at Metre Wavelengths.—R. Payne-Scott. (*Aust. J. sci. Res., Ser. A*, June 1949, Vol. 2, No. 2, pp. 228–231.) Observations were made of the enhanced radiation on August 5 and 6, 1948. The output on 60 and 85 Mc/s of a receiver having a bandwidth of 1.5 Mc/s was fed into a biased diode detector whose current/bias curve was measured. No difference was observed in the amplitude distribution of solar and thermal noise. Solar radiation does not reach us as a series of discrete frequencies separated at intervals of more than 2 Mc/s.

523.72 : 621.396.822] : 538.566.2 886

The Wave Equations for Electromagnetic Radiation in an Ionized Medium in a Magnetic Field.—Westfold. (*See* 976.)

523.854 : 621.396.822 887

The Position and Probable Identification of the Source of Galactic Radio-Frequency Radiation Taurus-A.—J. G. Bolton & G. J. Stanley. (*Aust. J. sci. Res., Ser. A*, June 1949, Vol. 2, No. 2, pp. 139–148.) The source was discovered by an interference method (412 of 1948), using a frequency of 100 Mc/s. It has a width of less than 6° and coincides with the Crab nebula—a former supernova.

551.510.535 888

Measurement of Height Distribution of Ionization in the Ionosphere.—G. Goubau. (*Arch. tech. Messen*, Nov. 1949, No. 166, pp. T99–T100.) Description of a method and apparatus used at Herzogstand for vertical-incidence measurements. Frequency ranges are 1–3 Mc/s and 3–9.5 Mc/s.

551.510.535 889

The Distribution of Ionization according to Height and the Recombination Coefficient of the Ionosphere F Layer.—A. A. Åynberg. (*Zh. eksp. teor. Fiz.*, June 1949, Vol. 19, No. 6, pp. 515–520. In Russian.) From experimental critical-frequency/height characteristics of the F layer the distribution of ionization n with height is determined. Two curves showing the daily variation of n at various heights from 230 to 350 km are plotted for July and November (Figs. 2a and 2b). A table is also given showing the day and night values of the recombination coefficient at various heights for the same two months. These values were calculated on the basis of the law of simple recombination. The results obtained in this paper are only approximate.

551.510.535 890

The Height of the F_2 Ionospheric Layer and the Relative Sunspot Number.—R. Eyfrig. (*Rev. sci., Paris*, Nov. 1948, Vol. 86, No. 3299, pp. 673–674.) The Huancayo monthly median values of the ratio m.u.f. to critical frequency for the F_2 layer, for a reference distance of 3000 km, are correlated with the corresponding Zurich mean values of the relative sunspot number for the period Jan. 1944 to April 1949. Both noon and midnight data are considered. The results show that the height of the F_2 layer increases during the increasing phase of the solar cycle and diminishes during the decreasing phase.

551.510.535 : 550.385 891

World Morphology of Ionospheric Storms.—E. V. Appleton & W. R. Piggott. (*Nature, Lond.*, 28th Jan. 1950, Vol. 165, No. 4187, pp. 130–131.) Superposed epoch analyses were made of noon f^0F_2 values during ionospheric storms. In temperate latitudes there is a marked rise in f^0F_2 for two days, followed by a sudden drop; recovery to normal takes several days. Near the equator, only the rise is found and this coincides with the drop at higher latitudes.

In temperate and high latitudes, sudden commencements are often accompanied by a rapid fall in f^0F_2 followed by a rise; this is often missed because it seldom lasts more than half an hour. Ionospheric storms are usually confined to a limited longitude range and are not equally developed in the north and south hemispheres.

551.510.535 : 621.3.087.4 892

Ionospheric Sounding Experiments in Germany.—W. Dieminger. (*Research, Lond.*, Dec. 1949, Vol. 2, No. 12, pp. 571–576.) A survey of recent experiments using high-power pulse transmitters (15–130 kW). The transmitters use master-oscillator and power-amplifier valves whose anodes are pulsed by a charged-line triggered-thyratron modulator. A variable-frequency sounder uses a servo motor to keep the transmitter in step with the receiver. For oblique-incidence work, transmitter pulse frequency and receiver timebase are controlled by crystal clocks to keep them in synchronism.

Experiments on the echoes scattered back to the transmitters via the E or the F regions and ground irregularities, and observations of the screening effect of the E region in oblique incidence transmission, preventing single-hop transmission via the F layer while permitting double-hop transmission, are described.

551.594.6 893

Study of the Wave-Forms of Atmospherics.—S. R. Khastgir & R. Roy. (*Phil. Mag.*, Nov. 1949, Vol. 40, No. 310, pp. 1129–1143.) Variations in the field strength of atmospherics were amplified in a wide-band amplifier (100–10 000 c/s) and applied to a c.r.o. with a linear timebase of duration normally between 1 and 10 ms. Drawings of the traces were made by hand. The waveforms were divided into eight types and their frequencies of occurrence determined. The main characteristics of these types are tabulated and discussed in relation to the physical properties of the lightning discharge.

LOCATION AND AIDS TO NAVIGATION

621.396.9 894

Surveillance Radar System at the Port of Le Havre.—R. Pélicant. (*Électronique, Paris*, Dec. 1949, No. 38, pp. 8–12.) A description of the shore-based installation which scans an area of 50-km radius around the port. The horizontal spreading of the beam precludes its use for precise navigation but it is in continual use at times of bad visibility for locating vessels and obstacles. Peak power is 180 kW; wavelength 10 cm; pulse duration 2μs with 1 000 pulses per sec; aerial height 45 m; angle of spread 3°; 4 different rotation speeds, the fastest being 10 r.p.m.

621.396.93 895

A Voltage Discriminator; its Application to Direction Discrimination.—J. Loeb, M. Jezo & C. Lombard. (*Ann. Télécommun.*, Feb. 1949, Vol. 4, No. 2, pp. 57–63.) A description of the principles, construction and performance of a v.h.f. d.f. device which has an angular sensitivity of 10^{-6} radian. The apparatus is basically a goniometer which compares two low-level h.f. voltages and gives a signal which actuates a motor in a direction determined by the sense of the input voltage. Signal-voltage difference sensitivity is of the order of 1 μV.

621.396.933 896

Crystal Control at 1 000 Megacycles for Aerial Navigation.—S. H. Dodington. (*Elect. Commun.*, Dec. 1949, Vol. 26, No. 4, pp. 272–278.) 1949 I.R.E. National Convention paper. A description of distance-measuring equipment, using the ground-beacon interrogation system, which is suitable for multiplex operation. Fifty-one 2-way channels are provided with a spacing of 2.4 Mc/s. Adjacent channel rejection is better than 70 db, while image rejection and all spurious responses within the distance-measuring band are at least 60 db down in both air and ground equipments. The ranging circuits provide an indication of distance up to 115 nautical miles; accuracy is within 0.2 mile.

629.13.052 : 621.396.933 897

Aircraft Radio Altimeters.—B. A. Sharpe. (*J. Inst. Nav.*, Jan. 1950, Vol. 3, No. 1, pp. 79–89.) A description of altimeters using (*a*) pulse radar systems, (*b*) f.m. of a c.w. carrier. Accuracy and sources of error are discussed. It is necessary to use both systems in order to cover the range 0–50 000 ft.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.5 898

Symposium on High Vacuum.—(*Industr. Engng Chem.*, May 1948, Vol. 40, No. 5, pp. 778–847 & 938–944.) Papers presented at a symposium arranged by the National Research Corporation and the American Chemical Society, Cambridge, Mass., 30th & 31st Oct. 1947. These include :—

- Development of High-Vacuum Technique.—S. Dushman.
- Design of High-Vacuum Engineered Plants.—H. C. Weingartner.
- Design of High-Vacuum Systems.—C. E. Normand.
- New Techniques in the Measurement of Pressures below 10 mm.—G. L. Mellen.
- Measurement and Control of Leakage in High-Vacuum Systems.—R. B. Jacobs.
- Measurement and Comparison of Pumping Speeds.—B. B. Dayton.
- Surface Phenomena useful in Vacuum Technique.—L. Apker.
- Eight papers on practical industrial applications of vacuum technique are also included.

535.312 : 546.212 : 621.396.11 899

Measurements of the Reflection Coefficient of Water at a Wavelength of 8.7 mm.—D. G. Kiely. (*Proc. phys. Soc.*, 1st Jan. 1950, Vol. 63, No. 361B, pp. 46–48.) An account is given of measurements (carried out in March 1947) of the reflection coefficient of water at a wavelength of 8.7 mm over a range of angles of incidence. The method employed is to measure the relative field strengths of the direct waves and waves reflected from a trough of water, using free-space propagation and high-gain aerials. The following electrical constants of water have been computed from the measured results (water temperature 11.1°C): refractive index 4.40 ± 0.24, dielectric constant 10.86 ± 2.21, absorption coefficient 2.91 ± 0.06, conductivity/frequency 12.82 ± 0.42, Brewster angle 79°.

535.312 : 546.212-16 : 621.396.11 900

Reflection Coefficient of Snow and Ice at V.H.F.—J. A. Saxton. (*Wireless Engr*, Jan. 1950, Vol. 27, No. 316, pp. 17–25.) A general formula is given for the reflection coefficient of plane waves incident on a layer of ice or snow on the earth's surface, taking account of the multiple reflections within the layer. The reflection coefficients of ice on sea water and of snow on land are calculated for frequencies of 300 and 3 000 Mc/s, for three angles of incidence, both for vertically and for horizontally polarized waves. The presence of such layers may materially affect the reflection coefficient and hence the vertical-coverage diagram of a v.h.f. transmitter.

537.228.1 : 546.431.82 901

The Piezoelectric Effect in Barium Titanate.—A. V. Rzhanov. (*Zh. eksp. teor. Fiz.*, June 1949, Vol. 19, No. 6, pp. 502–506. In Russian.) Experiments were conducted with polycrystalline samples of barium titanate and the piezo-effect was investigated under static conditions when pressure was applied to the sample, and also under dynamic conditions when longitudinal vibrations were excited. A number of experimental curves are plotted and the values are determined of the piezo-moduli corresponding to the application of the force in the direction of polarization and at right angles to it. The material is quite suitable for use in piezoelectric transducers.

538.22 902

On the Variation of Thermo-Remanent and Isothermally Remanent Magnetization of Fired Earths as a Function of the Applied Field.—J. Roquet. (*C. R. Acad. Sci., Paris*, 28th Nov. 1949, Vol. 229, No. 22, pp. 1135–1137.)

538.221 903

Magnetic Properties of Nickel-Cobalt and Related Alloys.—E. P. Wohlfarth. (*Phil. Mag.*, Nov. 1949, Vol. 40, No. 310, pp. 1095–1111.)

538.221 904

Investigation of the Magnetic After-Effect in Commercial Silicon- and Nickel-Iron Laminations.—H. Wilde. (*Frequenz*, Nov. & Dec. 1949, Vol. 3, Nos. 11 & 12, pp. 309–319 & 348–353.)

538.221 905

Some Ferromagnetic Properties of Mixed Ferrites of Nickel and Zinc.—C. Guilland & M. Roux. (*C. R. Acad. Sci., Paris*, 28th Nov. 1949, Vol. 229, No. 22, pp. 1133–1135.)

538.221 906

Macroscopic Magnetic Texture of Permalloy Tape.—I. Épelboim & A. Marais. (*C. R. Acad. Sci., Paris*, 28th Nov. 1949, Vol. 229, No. 22, pp. 1131–1133.)

538.221 : 061.3 907

Ferro-Magnetic Materials.—(*Elect. Times*, 10th Nov. 1949, Vol. 116, No. 3027, pp. 655–660.) A complete list of titles and authors of 20 papers read at a recent I.E.E. symposium, with brief summaries of 10 of these papers.

538.221 : 621.318.22 908

Structure and Properties of the Permanent-Magnet Alloys.—A. H. Geisler. (*Elect. Engng, N.Y.*, Jan. 1950, Vol. 69, No. 1, pp. 37–44.) Essentially the full text of a paper presented at the symposium on magnetics at the 1949 A.I.E.E. Winter Meeting, New York. The reactions during the formation of such alloys, and their properties during and after formation, are described.

621.315.59 909

The Mechanism of Electrical Conduction in Semiconductors, and their Technical Application.—K. Grosskurth. (*Fernmeldetech. Z.*, Jan. 1950, Vol. 3, No. 1, pp. 22–27.)

621.315.59 910

The Electrical Conductivity of Simple Semiconductors.—W. Ehrenberg. (*Proc. phys. Soc.*, 1st Jan. 1950, Vol. 63, No. 361A, pp. 75–76.) By combining an approximate expression for f_1 of the Busch-Labbart type (1820 of 1947) with Shifrin's results (2218 of 1946), an analytical expression for the conductivity of semiconductors valid for all temperatures becomes available, as well as a simple procedure for determining the constants. The expression is derived under certain limiting assumptions.

621.315.612.4 911

Alkaline-Earth Titanates as Dielectrics and a New Group of Piezoelectric Materials.—W. M. H. Schulze. (*Elektrotechnik, Berlin*, Dec. 1949, Vol. 3, No. 12, pp. 365–372.) A detailed discussion of the physical and electrical properties of these ceramics, with explanatory theory.

621.318.33 912

A Simply-Constructed Small Electromagnet of High Performance.—R. P. Hudson. (*J. sci. Instrum.*, Dec. 1949, Vol. 26, No. 12, pp. 401–404.) A Weiss-type electromagnet is described which will produce fields of 20 kilogauss in a pole gap of width 3 to 5 cm. The coils, designed to dissipate 100 kW, are layer wound with enamelled copper strip and encased in insulating material. Cooling is provided by circulating distilled water through a heat exchanger which is cooled by mains water. The coil cases are made watertight by means of rubber gaskets and can be dismantled easily for inspection.

668.3 : 621.396.611.21 913

Uses of a New Adhesive in Production of Piezoelectric Vibrators.—J.E.T. (*P.O. elect. Engrs' J.*, Jan. 1950, Vol. 42, Part 4, pp. 187–188.) An account of various applications of a thermosetting adhesive, Araldite, in the production of crystal vibrators. Araldite makes very strong joints between glass, ceramic and metal surfaces when cured at 160–240°C. It has been used (*a*) for attaching thin wires to crystal plates to act as supports and connections to electrodes, and (*b*) for cementing abrasive powders to the rim of disks used for cutting crystal slices. See also 3430 of 1949.

MATHEMATICS

517.63 : 517.432.1) : 621.3.015.3 914

The Laplace Transformation and the Study of Transient Phenomena.—S. Colombo. (*Ann. Télécommun.*, June & July 1949, Vol. 4, Nos. 6 & 7, pp. 210–222 & 233–249. Errata, Nos. 8/9, p. 306 & p. 328. Appendices, No. 10, pp. 358–362.) The first part of this article is concerned with generalities relating to functional analysis and to the mathematical study of transient electrical phenomena. Part 2 discusses the operational methods of Heaviside and the Bromwich-Wagner integral. In part 3 the theories of the symbolic calculus are outlined and their application to various problems of wave propagation is considered. The methods of the symbolic calculus are applied to the solution of integral equations and to the determination of certain asymptotic developments which occur in the mathematical treatment of transients in electrical circuits.

517.93 : 621.396.611.1 915

Harmonic and Subharmonic Response for the Duffing Equation.—Levenson. (*See* 847.)

681.142 916

Mathematical Machines.—H. M. Davis. (*Sci. Amer.*, April 1949, Vol. 180, No. 4, pp. 28–39.) A general discussion of the historical development of calculation machines, with some speculations about future developments.

681.142 917

The EDSAC — An Electronic Calculating Machine.—M. V. Wilkes & W. Renwick. (*J. sci. Instrum.*, Dec. 1949, Vol. 26, No. 12, pp. 385–391.) A large-scale electronic calculator which works in the binary system and uses ultrasonic delay units for the storage of orders and numbers. Punched tape is used for the input and a teleprinter for the output. The functions of the various units are described and the means by which orders are taken one by one from the store and executed are explained. See also 3448 of 1948, 1425, 2823 and 3458 of 1949.

681.142 918

Nonlinear Functions in an Analog Computer.—G. D. McCann, C. H. Wilts & B. N. Locanthi. (*Elect. Engng, N.Y.*, Jan. 1950, Vol. 69, No. 1, p. 26.) Summary of a paper presented at the 1949 A.I.E.E. Summer Meeting on "Application of the California Institute of Technology Computer to Nonlinear Mechanics and Servomechanisms". Two basic types of device have been developed for use with the computer. These are (*a*) resistance-switching Ge-diode circuits; (*b*) a c.r. tube/template/photocell servo system. Both devices can be used to represent general nonlinear impedances or amplification factors. Their application in different analyses is described. See also 3176 of 1949.

681.142 : 621-526 919

Electronics and Experimental Mathematics.—F. H. Raymond. (*Onde élect.*, Jan. 1950, Vol. 29, No. 274, pp. 30–44.) The basic principles of mathematical machines of the algebraic (analogue) and arithmetical types are briefly discussed. A general theory of linear servomechanisms with many variables is presented, their accuracy and stability are assessed and their application in the analogue type of computer is considered. The construction of computers of the type OME (Opérateur Mathématique Électronique), made by the Société d'Électronique et d'Automatisme, is based on servomechanism theory. These computers are designed for obtaining integro-differential equation solutions which satisfy either the conditions of Cauchy or given initial conditions. Nonlinear equations can be solved. The functional characteristics of the OME are described and its stability conditions are analysed.

517.53 : 621.392 920

Applications Physiques de la Transformation de Laplace. (Book Review)—M. Parodi. Publishers: Centre National de la Recherche Scientifique, Paris, 1948, 174 pp. (*Ann. Télécommun.*, March 1949, Vol. 4, No. 3, p. 104.) The first of a series of monographs on mathematical processes undertaken by the Centre of Mathematical Studies. It presents in condensed form the principal practical aspects of the Laplace transformation.

MEASUREMENTS AND TEST GEAR

531.76 921

Electrostatic Short-Time Measurement.—W. Schaaffs. (*Frequenz*, Oct. 1949, Vol. 3, No. 10, pp. 295–299.) Description of several simple capacitor methods for the measurement of times down to 0.5 ms.

531.764.5 : 621.396.91 922

On the Monitoring of the Rate of Quartz Clocks by means of Time Signals.—F. Conrad. (*Frequenz*, Sept. 1949, Vol. 3, No. 9, pp. 270–273.) Discussion of accuracy attainable. An example is given of records, for the month of September 1948, of time signals from Rugby (GBR) and Washington (WWV); the two curves are essentially parallel.

531.765 923

Millisecond Measurements in Research.—D. W. Gillings. (*Instrum. Practice*, May & June 1949, Vol. 3, Nos. 7 & 8, pp. 277–287 & 333–341.) A comparative survey of four basic types of millisecond chronograph is given, with details of the construction and performance of particular instruments. The recording units usually consist either of a moving-film camera which photographs the trace given by an ordinary oscillograph, or of a static camera in combination with an oscillograph giving a spiral trace. One non-recording type measures the potential of a standard capacitor charged at a known constant current during the time interval to be measured. Another uses an electronic counter to count the oscillations of a fixed-frequency oscillator during the required interval, and is particularly suitable for making routine measurements 'by the thousand'.

534.1 : 621.317.333.4 : 621.394'.396'.6 924

Fault Location in Transmission Equipment by Vibration Testing and Continuous Monitoring.—H. G. Myers. (*P.O. elect. Engrs' J.*, Jan. 1950, Vol. 42, Part 4, pp. 189–197.) The various types of fault in both transmitting and receiving equipment that can be readily discovered by means of suitable mechanical-vibration tests are discussed. Vibration test methods and apparatus are described and also continuous-monitoring methods which provide invaluable help to the maintenance engineer in solving recurrent-fault problems.

538.71 925

Three-Component Magnetometer.—J. W. Seaton. (*Electronics*, Jan. 1950, Vol. 23, No. 1, pp. 165–171.) A small saturable reactor is used for each component of the field. Each reactor is operated at 800 c/s, the magnitude and sense of the component being derived from observations of the amplitude and phase of the resulting 1 600 c/s voltage developed in the winding. A null method is used in which the steady magnetic component under observation is balanced out by the field derived from a coil carrying d.c. Accuracy of measurement is within ±0.1 milligauss.

621.3.011.3[083.74] : 621.3.016.35 926

The Stability of Inductance Standards.—G. H. Rayner & L. H. Ford. (*J. sci. Instrum.*, Jan. 1950, Vol. 27, No. 1, pp. 19–21.) Information acquired at the National Physical Laboratory since 1917 is reviewed. A brief description is given of the present working standard of mutual inductance, an oil-immersed type with a range of $-1\ \mu H$ to $11\,000\ \mu H$, whose value has remained constant to within ±2 parts in 10^5 during the past twelve years. This instrument, when used in a bridge, makes possible the certain detection of changes of 2 parts in 10^5 in the values of inductance or capacitance. The stability with time of substandard self-inductance and fixed mutual-inductance coils of various types is also discussed.

621.317.3 : 550.371 927

Apparatus for Measuring the Intensity of an Electric Field and its Applications.—I. M. Imyanitov. (*Zh. tekh. Fiz.*, Sept. 1949, Vol. 19, No. 9, pp. 1020–1031. In Russian.) Many papers have been published during the last two decades, describing various types of apparatus all utilizing the principle of the e.s. generator but designed to measure different electrical quantities such as current, voltage, charge, field strength etc. An attempt is made to generalize the theory of such apparatus, and a report is given together with a theoretical discussion of experiments with universal apparatus built by the author, in which an e.s. generator is used. The sensitivity of this apparatus is 0.01 V/cm per scale division, the area of the measuring plate 225 cm^2 and the minimum input impedance 15 MΩ. Various possible applications are discussed.

621.317.3 : 621.396.82 928

Electrical Noise. Experimental Correlation between Aural and Objective Parameters.—Maurice, Newell & Spencer. (*See* 989.)

621.317.3 : 621.396.822 929

A Broad-Band Microwave Noise Source.—W. W. Mumford. (*Bell Syst. tech. J.*, Oct. 1949, Vol. 28, No. 4, pp. 608–618.) The theory of measurement of receiver noise figure is discussed; an ordinary resistive noise source is unsuitable for accurate microwave measurements because it cannot be raised to a sufficiently high temperature. An alternative source is an electrical gas discharge, and investigation shows that a commercial fluorescent lamp is satisfactory. At 4 000 Mc/s its effective temperature is about 11 000 °K, which is convenient for measuring noise figures of 20 db or less. The noise power is practically independent of the fluorescent coating and the current density, and is only slightly affected by the room temperature. The lamp lends itself readily to broad-band impedance matching in a waveguide.

621.317.3 : 621.396.822 : 621.396.645 930

Study of Some Experimental Methods used for Measurement of Noise of Centimetre-Wave Amplifiers.—M. Denis. (*Ann. Radioélect.*, Jan. 1950, Vol. 5, No. 19, pp. 27–35.) Noise factor and signal-noise ratio are defined. The measurement system considered comprises a crystal mixer to superimpose u.h.f. on a local oscillator signal of several Mc/s, an amplifier with a relatively very narrow bandwidth at this latter frequency, and a quadratic detector (crystal or thermocouple). The effect of amplifier and detector on the value found for the noise factor is discussed. The amplifier and mixer circuits are described. The measurement of signal-noise ratio where the output signal is large is then considered. The method consists of converting the noise frequency to one at which comparison can be made with a saturated diode. The methods used apply specially to measurements on travelling-wave valves; a report on experimental results will be published later.

621.317.3.029.64 931

Dynamical Measurement of Q-Factors and Natural Frequencies of Cavity Resonators with a Single Coupling Element.—M. Denis & S. Couybes. (*Ann. Radioélect.*, Jan. 1950, Vol. 5, No. 19, pp. 54–61.) A theoretical analysis and a description of the measurement equipment used. A reflex klystron source has a 16-db attenuator in an extension of the output line. Such an arrangement acts as if a source of e.m.f. were located at the output end of the attenuator. A detector loosely coupled to the attenuator input follows the power variations as a function of frequency. A transformer section couples the artificial source to the resonator. In this section is a second quadratic detector. Measurements are made by

comparison of the two detector outputs on an oscillograph screen. The method is especially useful for measuring high-Q circuits and affords a quick means of obtaining the characteristics of a rhumbatron loaded by an electron beam, under varying circuit conditions. See also 1076 of 1948.

621.317.331 **932**

An Absolute Measurement of Resistance by the Wenner Method.—J. L. Thomas, C. Peterson, I. L. Cooter & F. R. Kotter. (*Bur. Stand. J. Res.*, Oct. 1949, Vol. 43, No. 4, pp. 291–353.) A full report of the apparatus and method developed for the measurement of resistance in terms of length, time, and the permeability of free space. A mutual inductor was constructed and its inductance determined from its dimensions to within a few parts in a million. Using Wenner's commutated d.c. method, the value found for the unit now maintained at the Bureau was 0.999 994 absolute ohm.

621.317.336 + 621.396.67 **933**

Antennas and Open-Wire Lines: Part 3 — Image-Line Measurements.—Conley. (*See* 810.)

621.317.336 **934**

Methods for measuring Impedances of Circuits shunted with a Negative Resistance.—Kh. I. Cherne. (*Radiotekhnika, Moscow*, Sept./Oct. 1949, Vol. 4, No. 5, pp. 63–70. In Russian.)

621.317.336 : 621.315.6.029.6 **935**

Abac of the Function $\tanh z/z$ for the study of Dielectrics at U.H.F.—J. Benoit. (*Ann. Télécommun.*, Jan. 1949, Vol. 4, No. 1, pp. 27–32.) This function of the complex number $z = u + jzw$ occurs in measurements of the impedance of a section of dielectric inserted at the end of a short-circuited line. The method of construction of the abac, which relates measured impedance with dielectric characteristics ϵ' and ϵ'', is described and geometric properties are established which greatly simplify the plotting of the curves for $u =$ C and $w =$ C.

621.317.35 **936**

Frequency Analyser with Rapid Automatic Scanning.—L. Pimonov. (*Ann. Télécommun.*, July 1949, Vol. 4, No. 7, pp. 257–272.) The principal automatic analysis systems are reviewed. The principle of sound analysis by the heterodyne method is illustrated by a detailed description of an analyser comprising (*a*) input amplifier, (*b*) triode-hexode phase converter, (*c*) compensated modulator, (*d*) high-Q oscillator, (*e*) quartz filter, (*f*) measuring device. The relation of speed of analysis to selectivity is considered and the analysis of ultrasonic frequencies is discussed. Different applications of the analyser are considered and records of different sound spectra are reproduced.

621.317.7 : 621.396.813 **937**

Study and Construction of a Distortion Meter.—F. Haas. (*Toute la Radio*, Feb. 1950, Vol. 17, No. 143, pp. 87–91.) A band-stop filter is interposed between an amplifier and a valve voltmeter to cut out the fundamental frequency. The filter consists of a resonance bridge between two attenuators. Measurement is made either on the valve voltmeter or on an oscillograph. Full circuit details, with component values, are given.

621.317.723 : 621.317.733 **938**

The Electrometer as Bridge (balance) Indicator.—T. Gast. (*Frequenz*, Sept. 1949, Vol. 3, No. 9, pp. 264–270.) Detailed discussion of the use of e.s. instruments as null indicators. Advantages are that they take no power, are practically independent of frequency and can be readily adapted for remote indication or for use in self-balancing bridges.

621.317.725.024.2 **939**

A Single-Pulse Voltmeter.—G. T. Rado, M. H. Johnson & M. Maloof. (*Rev. sci. Instrum.*, Dec. 1949, Vol. 20, No. 12, pp. 927–929.) The instrument consists of a gated cathode-follower bridge in conjunction with a ballistic indicator. It measures the average amplitude of a single pulse over an arbitrarily selected interval of 2 ms, with a random error of 0.2% of full-scale deflection.

621.317.73 **940**

Impedance Measurements with Directional Couplers and Supplementary Voltage Probe.—B. Parzen. (*Elect. Commun.*, Dec. 1949, Vol. 26, No. 4, pp. 338–343.) Reprint. See 155 of January.

621.317.733 **941**

A New Design of Wheatstone Bridge.—W. L. Surman. (*P.O. elect. Engrs' J.*, Jan. 1950, Vol. 42, Part 4, pp. 209–212.) Some disadvantages of the Post-Office-box type of bridge have been eliminated in the No. 2 bridge, which is self contained, measures resistances from 1 to 9 999 Ω and has an accuracy of ±0.5 to ±1%. A commercial version, No. 3, has P.O. box accuracy of ±0.2%. Four decade resistor combinations are included in the comparison arm; operation of the range switch gives a visual indication of the position of the decimal point.

621.317.733 **942**

New Bridge Technique.—T. Roddam. (*Wireless World*, Jan. 1950, Vol. 56, No. 1, pp. 8–10.) Comment on 2271 of 1949 (Yates). Square-wave excitation, instead of sine-wave, can be used for measuring complex impedances, the bridge output being applied to a c.r.o. The waveform obtained gives an indication of the type of impedance being measured and so facilitates balancing.

621.317.733 **943**

The Impedance Bridge (No. 2135) constructed by the S.A.C.M. for the Service of L.S.G.D. [Lignes Souterraines à Grande Distance].—G. Chardon & R. Blondé. (*Câbles & Transmission, Paris*, Jan. 1950, Vol. 4, No. 1, pp. 58–65.) The construction of this portable bridge is described and performance figures are given. Its chief features are a series arrangement of resistance and capacitance standards for the measurement of impedance, and plug-in resistor and capacitor units for measurement of zero and positive reactance. A similar system converts the instrument for measuring admittance unbalance. Precision measurements can be made up to 10 000 Ω between 200 c/s and 10 kc/s and up to 400 Ω at 300 kc/s.

621.317.733 **944**

Capacitance Bridge with Mechanical Rectifier and Mirror Galvanometer. Elimination of Errors due to Harmonics.—F. Koppelmann. (*Frequenz*, Sept. 1949, Vol. 3, No. 9, pp. 259–264.) An account of improvements of the bridge previously described (3181 of 1948) and investigation of sources of error. Adjustment of the timing of the rectifier contacts enables different harmonics of the supply to be cut out. See also 1429 and 3464 of 1949.

621.317.763 : 621.392.26 **945**

Interferometer for Hertzian Microwaves.—T. Kahan. (*J. Phys. Radium*, June 1947, Vol. 8, No. 6, p. 192.) Description of apparatus used to measure wavelengths in the 3-cm band. Two sections of rectangular waveguide (30 × 10 mm) are soldered together in the form of a cross to make the four arms of the instrument. The ends, A and B, of two opposite arms are fitted with horns to concentrate the linearly polarized beams. The klystron generator output is fed into A. A thin trolitul plate fitted with parallel strands of wire is fixed inside the waveguides diagonally across the junction of the four

arms. This performs the same function as the semi-transparent plate in Michelson's interferometer. By means of an adjustable reflector R_1 facing the end B and a fixed, grid-type reflector R_2 at the end of the third arm, the two reflected waves are superimposed in the fourth arm, which is terminated with a piston wave trap. Here a crystal detector, which is coupled to an oscilloscope, can be adjusted in position along a fine longitudinal slot. By modulating the generator input at 1 000 c/s and controlling the wave-train by adjustment of R_1, the wavelength may be measured on the oscilloscope. The reflection coefficient of the disk R_2 and terminal impedance may also be measured.

621.396.615 946

The New General-Radio Generator 1001-A.—A. G. Bousquet. (*Toute la Radio*, Feb. 1950, Vol. 17, No. 143, pp. 93–96.) A detailed description of the construction and performance of an instrument which replaces the former standard model No. 605. It covers from 5 kc/s to 50 Mc/s in 8 ranges. The output may be modulated in amplitude up to 80% either by an external source with a frequency range from 20 c/s to 15 kc/s, or by a 400-c/s internal source. Output jacks provide (*a*) a voltage continuously variable between 0.05 μV and 200 mV or (*b*) a fixed voltage of 2 V. A complete circuit diagram giving all component values is shown.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

536.5 : 621.396.822 947

An Absolute Noise Thermometer for High Temperatures and High Pressures.—J. B. Garrison & A. W. Lawson. (*Rev. sci. Instrum.*, Nov. 1949, Vol. 20, No. 11, pp. 785–794.) A null device for determining the ratio of two absolute temperatures by balancing the thermal noise voltages developed across two resistors at the relevant temperatures. Temperatures up to 1 200°K have been measured with an accuracy of 0.1%, with an observation time of 2 minutes. Calibration of the thermometer elements is independent of the chemical composition and past physical history of the resistance material used and of the absolute pressure.

539.16.05 948

Dead Times of Self-Quenching Counters.—B. Collinge. (*Proc. phys. Soc.*, 1st Jan. 1950, Vol. 63, No. 361B, pp. 15–24.)

551.508.11 : 621.396.91 949

The Swiss Radiosonde.—J. Lugeon, P. Ackermann & M. Bohnenblust. (*Annales de la Station centrale suisse de Météorologie*, 1948, 17 pp. In French. Reprint.) Illustrated description of the different components and discussion of their operation. The bimetallic-spiral thermometer, aneroid barometer and goldbeaters'-skin hygrometer are mechanically coupled to pointers carrying at their ends small capacitor plates whose location is scanned by a similar plate carried on a clockwork-driven arm rotating once every 30 sec. This arrangement is used for f.m. of a small 100-Mc/s transmitter. The equipment used for receiving and recording the transmitted signals is also described.

551.508.5 : 621.317.755 950

Acoustic Anemometer-Anemoscope.—R. E. Corby. (*Electronics*, Jan. 1950, Vol. 23, No. 1, pp. 88–90.) An instrument giving an instantaneous display of wind direction and velocity on a c.r.o. screen. 60-c/s pulses from an acoustic transmitter are received by transducers equidistant from the transmitter in the four cardinal directions. A discriminator analyses the Doppler effect of the wind on the time of arrival of the signals at the four receivers, and actuates an indicator. Wind speed is given by the length of the radial trace on the c.r.o. screen.

621.318.572 : 531.77 951

Precision Measurement of Rotary Motion.—H. J. Finden. (*Electronic Engng*, Jan. 1950, Vol. 22, No. 263, pp. 2–8.) The measurement is carried out by obtaining a pulsed signal whose pulse repetition rate is directly proportional to rotation speed, and then electronically counting the number of pulses per second, the time scale being derived from a reference standard. Circuits of frequency dividers, gate relays, amplifiers and counters are given for equipment designed to give visual indication of r.p.m. for speeds varying from 500 to 20 000 r.p.m. A counter sufficiently light and compact for flight test work can be constructed to read to the nearest 10 r.p.m. over a range of 4 000–20 000 r.p.m.

621.365.54† : 621.314.57 952

A New Method for Converting D.C. Energy into High-Frequency A.C. Energy with a High Efficiency, specially intended for High-Frequency Induction Heating.—T. Douma. (*Commun. News*, June & Oct. 1949, Vol. 10, Nos. 2 & 3, pp. 52–68 & 69–82.) The method consists essentially in charging the capacitor of a resonant circuit from a d.c. source through a special type of Hg-vapour valve. During a controllable period following the charging process, an oscillatory discharge can take place through a load coil. The experimental Hg-discharge tubes are semicircular in form, with ends shaped like large flat dishes to accommodate the pools of Hg used for anode and cathode. A capacitive igniter projects through the cathode pool and positive h.v. pulses are applied to it in order to render the valve conducting. Detailed mathematical theory of the operation of such a device is given. An efficiency of 99% is possible at frequencies up to about 10 kc/s. Operational difficulties are briefly discussed. An output of 10 kW (4 A at 2 500 V) has been reached with a relatively small valve, but after 60 hours operation at this loading, operation became unsatisfactory. With a larger valve and better cooling arrangements, an output of 40 kW has been obtained for 30 hours, an output of 80 kW being reached occasionally. Valves or combinations of valves are aimed at which can handle continuously an output of 100 to 300 kW.

621.365.54† : 621.785.6 953

Automatic Induction Hardening speeds Dodge Output.—C. H. Wick. (*Machinery, N.Y.*, Nov. 1949, Vol. 56, No. 3, pp. 194–198.) Applications of induction heating in selective surface hardening of countershaft cams, resulting in faster heating and reduced distortion and scaling.

621.365.55† 954

The Physical Principles of Electrical Dielectric Heating of Nonconducting Materials.—F. Walter. (*Frequenz*, Oct. 1949, Vol. 3, No. 10, pp. 299–306.)

621.38 955

Papers Summarized for Conference on Electronic Instrumentation.—(*Elect. Engng, N.Y.*, Jan. 1950, Vol. 69, No. 1, pp. 68–71.) Authors' summaries of the following papers presented at the joint A.I.E.E. and I.R.E. Conference on Electronic Instrumentation in Nucleonics and Medicine, New York, 1949:

Low-Frequency Spectrography: Some Applications in Physiological Research.—R. R. Riesz.

Electrical Methods of Blood-Pressure Recording.—F. Noble.

A Stable D.C. Amplifier for Biological Recording.—H. Grundfest.

A 25-Channel Recorder for Mapping the Electric Potential Gradients of the Cerebral Cortex: Electroiconograms.—J. C. Lilly.

Medical Application of Ionizing Radiations.—E. H. Quimby.

An Automatic Isodose Recorder.—G. J. Hine.

The Measurement of Low-Energy Beta Ray Emitters: Ionization Chamber Techniques.—N. A. Bailey.

Proportional Counters for the Measurement of Soft Radiation.—C. J. Borkowski.

Small Geiger Counters for Biological Applications.—C. V. Robinson.

Phosphors for Scintillation Counters.—R. H. Gillette. Solids for Radiation Detection.—R. M. Lichtenstein.

Some Design Features of Electrical Counting Systems.—N. F. Moody.

Desirable Improvements in Nucleonics Instrumentation.—J. B. H. Kuper.

Criteria in the Selection of Radioisotopes for Industrial Use.—E. T. Clarke.

621.38.001.8 956

Electronic Instruments in Research and Industry.—(*Nature, Lond.*, 4th Feb. 1950, Vol. 165, No. 4188, pp. 181–182.) Report of the second symposium arranged by the Electronics Group of the Scientific Instrument Manufacturers' Association, London, November 1949, with short summaries of the papers presented. See also 957 below.

621.38.001.8 957

Electronic Equipment.—(*Electronic Engng*, Jan. 1950, Vol. 22, No. 263, pp. 34–35.) Descriptions of a selection of equipment exhibited at the Scientific Instrument Manufacturers' Association Symposium, London, November, 1949.

621.38.001.8 : 621.791.7 958

Some Particular Applications of Electronic Circuits to the Control of Resistance-Welding Machines.—E. Thurion. (*Bull. Soc. franç. Élect.*, Jan. 1950, Vol. 10, No. 100, pp. 47–53.) A review of some recent developments in precision welding technique.

621.383 : 621.365.54† : 621.791.052 959

Phototube Controls R.F. Welding.—H. H. Wittenberg. (*Electronics*, Jan. 1950, Vol. 23, No. 1, pp. 91–93.) Precise automatic control of the welding of small parts is provided by photocell equipment operated by the radiation from the glowing weld. The generator is switched off half a second after the welding metal flows.

621.384.611.2† 960

On the Capture of Particles into Synchrotron Orbits.—T. R. Kaiser. (*Proc. phys. Soc.*, 1st Jan. 1950, Vol. 63, No. 361A, pp. 52–66.) Theory is given for the case in which the r.f. accelerating voltage rises from zero to its maximum value in a finite time. If this time is long enough, all particles originally within a band of width $1/\sqrt{2}$ times the maximum width of the final stable region for synchrotron phase oscillations will be captured, irrespective of the initial phase at which they enter the accelerating gap. This conclusion appears to be in agreement with the observed characteristics of existing electron synchrotrons.

621.384.611.2† 961

Experiments on Electron Capture and Phase Stability in a 14-MeV Synchrotron.—T. R. Kaiser & J. L. Tuck. (*Proc. phys. Soc.*, 1st Jan. 1950, Vol. 63, No. 361A, pp. 67–74.) Observations were made of the effects on output of interrupting the r.f. accelerating voltage for definite intervals during the acceleration process. Capture is more efficient than expected and for short interruption periods the particle loss is negligible. The results are in quantitative agreement with Kaiser's theory (960 above).

621.384.612.† 962

Operation of the 184-in. Cyclotron.—L. R. Henrich, D. C. Sewell & J. Vale. (*Rev. sci. Instrum.*, Dec. 1949, Vol. 20, No. 12, pp. 887–898.) The operation is reviewed in terms of the theory developed by Bohm & Foldy (1542 of 1947 and 793 of 1948). Certain relevant data on the properties of the magnet and rotating capacitor are presented.

621.385.38.001.8 963

Examples of the Industrial Application of Gas-Filled Valves.—R. Kretzmann. (*Z. Ver. dtsch. Ing.*, 15th Sept. 1949, Vol. 91, No. 18, pp. 457–462.) Specific applications of thyratrons and relay valves for control purposes. An extensive bibliography is given.

621.385.833 964

An Electron Gun for an Electron Microscope.—V. N. Vertsner. (*C. R. Acad. Sci. U.R.S.S.*, 11th Aug. 1947, Vol. 57, No. 5, pp. 459–461. In Russian.) Description of a gun developed at the State Optical Institute. The main feature of the gun is that the filament can be accurately displaced, horizontally as well as vertically, with respect to the diaphragm aperture while the microscope is in use.

621.385.833 965

Electron Optics of the Three-Stage Electron Microscope.—C. E. Challice. (*Proc. phys. Soc.*, 1st Jan. 1950, Vol. 63, No. 361B, pp. 59–61.) Calculations by Gaussian optical methods, with ray diagrams deduced from the tabulated numerical results.

621.385.833 966

An Improved Method of Numerical Ray Tracing through Electron Lenses.—G. Liebmann. (*Proc. phys. Soc.*, 1st Dec. 1949, Vol. 62, No. 360B, pp. 753–772.) The general ray equation is integrated with the help of Taylor's series, including terms up to those of the fourth order. The method is first developed for combined e.s. and e.m. lenses and then specialized for lenses of pure e.s. or pure e.m. type. Space-charge effects within the lenses are also considered.

621.385.833 967

Electron-Optical Observation of Magnetic Fields.—L. Marton & S. H. Lachenbruch. (*Bur. Stand. J. Res.*, Oct. 1949, Vol. 43, No. 4, pp. 409–428.) A description of the electron optical analogue of the schlieren method and a related shadow method. See also 199 of January.

621.396.615 : 534-6 968

An Ultra-Low-Frequency Oscillator.—J. E. Stone. (*Electronics*, Jan. 1950, Vol. 23, No. 1, pp. 94–95.) The thermal lag of a current-carrying thermistor enables it to be used as an inductance in the resonant circuit of a subsonic oscillator. An approximately sinusoidal waveform can be obtained over a frequency range of 0.1–0.02 c/s.

621-57 : 621.318 969

Magnetic Fluid Clutch.—K. E. Wakefield. (*Gen. elect. Rev.*, Dec. 1949, Vol. 52, No. 12, pp. 39–43.) An account of the results of investigations with special reference to high-duty devices capable of handling several thousand horsepower. See also 50 of January (Parker) and 237 of January (Nelson).

621.791.3 : 534.321.9 970

Ultrasonic Soldering Iron for Light Alloys.—(*Engineer, Lond.*, 23rd Dec. 1949, Vol. 188, No. 4900, p. 737.) A removable copper bit is mounted in front of a magnetostriction transducer excited from a separate generator. Ultrasonic vibrations at about 20 kc/s destroy the oxide film on the working surface. No flux is needed.

621.791.352 : 621.365.54† 971

Industrial Brazing by Pulse Techniques.—J. L. Reinartz. (*Electronics*, Jan. 1950, Vol. 23, No. 1, pp. 78–80.) Extremely high values of peak r.f. power are applied in short-duration pulses in order to reduce the heating, by thermal conduction, of parts near to the joint being brazed. The method was developed for valve manufacture but may have many other applications.

629.13.055.2 972

A Shielded Hot-Wire Anemometer for Low Speeds.—L. F. G. Simmons. (*J. sci. Instrum.*, Dec. 1949, Vol. 26, No. 12, pp. 407–411.) "A low-speed anemometer is described which comprises an electrically heated wire and a thermocouple contained in a double-bore silica tube. In an air current the temperature change caused by heat loss is registered by the thermocouple and, when the instrument is suitably calibrated, the e.m.f. developed serves as a measure of the speed of flow."

621.365.54/.55† 973

Radio-Frequency Heating. [Book Review]—L. Hartshorn. Publishers: Allen & Unwin, London, 237 pp., 21s. (*Wireless Engr*, Jan. 1950, Vol. 27, No. 316, p. 31.) "This book covers both induction and dielectric heating and is concerned mainly with the principles of r.f. heating . . . A good point about the book is that it deals mainly with just those points upon which general information is lacking—the heating coil or electrode and the material itself. . . . Each chapter contains an extensive bibliography and the book undoubtedly forms a useful contribution to the literature."

621.38.001.8 974

Basic Electronics. [Book Review]—R. G. Kloeffler & M. W. Horrell. Publishers: J. Wiley & Sons, New York, and Chapman & Hall, London, 1949, 435 pp., $5.00. (*Electronics*, Jan. 1950, Vol. 23, No. 1, pp. 211–212.) "Coverage of the field of electronics from high-powered industrial circuits to low-level communications circuits is complete and comprehensive without being sketchy. Liberal references to the literature suggest further study for those whose interests or needs go beyond the scope of the general text."

PROPAGATION OF WAVES

538.566 975

The Normal-Wave Method and its Application to Multi-Layer Plane Media.—P. E. Krasnushkin. (*C. R. Acad. Sci. U.R.S.S.*, 1st Sept. 1947, Vol. 56, No. 7, pp. 687–690. In Russian.) A mathematical investigation of wave processes due to given currents in an unlimited, isotropic, non-magnetic ideal medium, the dielectric constant of which depends only on one Cartesian co-ordinate.

538.566.2 : 523.72 : 621.396.822 976

The Wave Equations for Electromagnetic Radiation in an Ionized Medium in a Magnetic Field.—K. C. Westfold. (*Aust. J. sci. Res., Ser. A.*, June 1949, Vol. 2, No. 2, pp. 169–183.) In the atmospheres of the sun and earth, the refractive index is, in general, a function of the angle between the direction of propagation and the magnetic field, the intensity of this field, electron density, and frequency. Surfaces of constant refractive index cannot be realized and the classical theory is inadequate.

The new treatment is based on axes parallel to the principal directions of the medium. A new form of the Appleton-Hartree equation for complex refractive index is obtained. Three possible types of plasma oscillation and their application to solar noise bursts are discussed. See also 884 and 885 (Payne-Scott).

538.566.2 : 621.396.11 977

Propagation of Electromagnetic Waves in a Stratified Medium.—G. Eckart. (*Ann. Télécommun.*, May & June 1949, Vol. 4, Nos. 5 & 6, pp. 142–154 & 223–230.) A mathematical treatment for propagation in a medium whose dielectric constant ϵ is a linear function of the z coordinate. The case of waves which transform into plane waves if the slope (α) of the ϵ curve tends towards zero is particularly considered, and also that of a medium consisting of two or three layers with different values of α. In the latter case reflection occurs at the discontinuities. Numerical calculations are made.

A solution for the general case, when the values of α are small, is developed in powers of α, and the coefficients of reflection and transmission are determined. Both horizontal and vertical polarization are discussed. Theory is also given of the Zenneck surface wave in a stratified medium.

621.396.11 978

Coupling of the Ordinary and Extraordinary Rays in the Ionosphere.—T. L. Eckersley. (*Proc. phys. Soc.*, 1st Jan. 1950, Vol. 63, No. 361B, pp. 49–58.) An approximate solution to the wave equation is given for propagation in an ionosphere in which the gradient of the density N is in the vertical, z, direction only, and in which account is taken of the earth's magnetic field. It corresponds exactly to the ray theory and expresses a quantity Z, which is the z derivative of the phase function S, by a quartic equation. Z can be represented as a function of ζ (which is proportional to N) on a four-sheeted Riemann surface, and the branch points are studied for the case of vertical incidence for which Z becomes the refractive index. By considering the branch points in the complex ζ plane, the amount of the coupling between the ordinary and the isolated extraordinary branches of the (Z, ζ) curves can be expressed as a function of the obliquity of the magnetic field. The triple splitting of rays reflected from the ionosphere, observable where the field is nearly vertical, can thus be explained, and the theory is substantiated by the observation that the polarizations of the echoes on the (P, f) records are ordinary, ordinary and extraordinary in order of increasing critical frequencies, as given by the branches of the (Z, ζ) curves.

621.396.11 : 535.312 : 546.212 979

Measurements of the Reflection Coefficient of Water at a Wavelength of 8.7 mm.—Kiely. (*See* 899.)

621.396.11 : 535.312 : 546.212-16 980

Reflection Coefficient of Snow and Ice at V.H.F.—Saxton. (*See* 900.)

621.396.11 : 550.837.7 981

Propagation of Electromagnetic Waves in Earth.—O. C. Haycock, E. C. Madsen & S. R. Hurst. (*Geophys.*, April 1949, Vol. 14, No. 2, pp. 162–171.) Consideration of attenuation, present-day transmitter technique, and aerial-length limitations, indicates that frequencies in the range 300–1 000 kc/s are the most practical for propagation through considerable thicknesses of ground.

621.396.11 : 621.396.8].029.6 982

The Propagation of Ultra-Short (Quasi-Optical) Waves.—W. Lehfeldt. (*Arch. elekt. Übertragung*, July–Dec. 1949, Vol. 3, Nos. 4–9, pp. 137–142, 183–186, 221–228, 265–269, 305–312, 339–346.) A comprehensive review including both theory and experimental results. Part 1 discusses (*a*) s.w. propagation characteristics ($\lambda < 10$ m); (*b*) refraction in a homogeneous troposphere and calculation of range; (*c*) discontinuities in refractive index causing reflection, and fading or extreme ranges of s.w. signals; (*d*) steep refraction gradients leading to total reflection; (*e*) an estimation of the limiting wave-

length (about 5 m) for sky-wave propagation: effects due to ion concentration; (f) scattering and absorption by water drops.

Part 2 deals with propagation measurements for wavelengths from about 6 m to 9 cm. The experiments were conducted in both hilly and flat country and also over the Baltic Sea. Equipment and measurement technique are described. A special section discusses the statistical evaluation of the distribution of atmospherics. Abnormal ranges are noted. Theoretical and experimental results are compared.

621.396.812.029.63 983

Interference Fading in the Decimetre-Wave Band caused by Humidity and Temperature Changes in the Lower Layers of the Atmosphere.—A. Grün & W. Kleinsteuber. (*Arch. elekt. Übertragung*, Sept. 1949, Vol. 3, No. 6, pp. 209–219.) Observed interference phenomena are explained in terms of (a) the bending of rays caused by gradients of the refractive index due to weather conditions, and (b) the interaction between the direct ray and the ray reflected from the earth. Consideration of the path difference of the two rays confirms that field strength variations may be calculated from the known behaviour of the refractive index. Graphical methods of calculation are described. Mean values and variations of the refractive-index gradients derived from observations of received signals are in good agreement with meteorological observations. A wavelength of about 50 cm appears especially advantageous for communication. The possible application of such observations for meteorological purposes is indicated.

621.396.97 : 621.396.619.13 : 621.396.8 984

Propagation Measurements near Geneva for Experimental F.M. Broadcasting Transmissions on 93 Mc/s.—W. Ebert. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st Oct. 1949, Vol. 27, No. 5, pp. 209–223. In German.) The Philips equipment used for these tests is described; circuit diagrams of the Type-FZ115a transmitter and Type-FM46 receiver are given. During the measurements the 750-W transmitter was operated at about half power. Field strength and quality of reception are recorded for 40 locations within a radius of 60 miles from the transmitter, (a) using a $\lambda/2$ dipole with vertical polarization, (b) using a horizontally polarized turnstile array. The latter gave appreciably better results, the field strength at any particular point being approximately 2.5 times that obtained with a vertical dipole and the same transmitter power. A further report is to be given later.

621.396.11 985

Radio Research Special Report No. 19. Lateral Deviation of Radio Waves reflected at the Ionosphere. [Book Notice]—W. Ross. Publishers: H. M. Stationery Office, London, 9d. (*Gost Publ., Lond.*, Nov. 1949, p. 19.)

RECEPTION

621.396.621 : 621.385 986

The Sensitivity of Receiving Valves.—H. Rothe. (*Arch. elekt. Übertragung*, Oct. 1949, Vol. 3, No. 7, pp. 233–240.) The noise factor which limits receiver sensitivity is calculated for grounded-cathode, grounded-grid and grounded-anode amplifier circuits. Neglecting the reactive effect of the output circuit, all three types of circuit have the same noise factor. The characteristics of various valves are compared and discussed with reference to operation at high frequencies. The grounded-grid triode circuit, on account of its lower equivalent noise resistance, gives at v.h.f. a greater sensitivity than the grounded-cathode pentode circuit.

621.396.621.029.62 : 621.396.933 987

Airways V.H.F. Communications Receiver.—A. H. Wulfsberg. (*Electronics*, Jan. 1950, Vol. 23, No. 1, pp. 81–83.) A receiver covering the range 108–136 Mc/s and using the double superheterodyne principle has a 100-db image-rejection ratio and an 80-db attenuation of other spurious responses. Series and shunt noise limiters are used in a special noise-balancing circuit, full details of which are given. Both the first and second local-oscillator frequencies are crystal-controlled. Output is 1W into a 600-Ω load.

621.396.621.5 988

Methods of Calculating the Sensitivity of Mixer Stages for Decimetre and Centimetre Waves.—Mataré. (*See* 855.)

621.396.82 : 621.317.3 989

Electrical Noise. Experimental Correlation between Aural and Objective Parameters.—D. Maurice, G. F. Newell & J. G. Spencer. (*Wireless Engr*, Jan. 1950, Vol. 27, No. 316, pp. 2–12.) The dependence of the amount of annoyance, defined as the disturbance caused by noise interference in the presence of a broadcast programme, upon the objective parameters of the noise was studied in order to devise a suitable measuring instrument which will simulate a listener's ear. The significant parameters are the receiver bandwidth and the pulse repetition frequency, considering the noise as repeated Heaviside unit impulses and the effect of the peak value on the annoyance to be logarithmic. Random fluctuation noise in the band 250 c/s to 10 kc/s was used as a standard; a number of observers compared this with seventeen different types of test noise under various combinations of the parameters. Simultaneously, the noise levels were recorded on six different types of noise meters preceded by aural-weighting networks. The results showed that a meter indicating the mean square value of the input waveform less the mean value was the most suitable of those available for the subjective measurement of the interference; suggestions for its improvement are given.

621.396.82 : 621.396.619.13 990

Investigation of Interference Phenomena in the Region between Two F.M. Common-Wave Transmitters.—H. Fricke, L. Pungs & K. H. Schmitter. (*Frequenz*, Oct. 1949, Vol. 3, No. 10, pp. 277–289.) The output voltage curve of a receiver situated between two f.m. transmitters operating on the same carrier frequency exhibits pronounced peaks which are due to h.f. and l.f. phase shifts. The number of these peaks within a given frequency band serves as a measure of the interference to programme reception. Harmonic analysis indicates that with increasing number of such peaks the energy distribution trends more and more towards the higher frequencies, so that by using a suitable low-pass filter the interference can be almost completely eliminated. Measurements on model equipment are described which confirm the theoretical conclusions.

621.396.62 + 621.396.662.3 + 621.396.645 991

Applications of the Electronic Valve in Receiving Sets and Amplifiers. [Book Review]—B. G. Dammers, J. Haantjes, J. Otte & H. van Suchtelen. Publishers: Philips Telecommunication Industries, Hilversum, Holland, 467 pp. (*Commun. News*, Oct. 1949, Vol. 10, No. 3, pp. 98–100.) The first volume of a series of three books on this subject. Only a.m. is considered. The various sections deal with (a) r.f. and i.f. amplification, (b) frequency changing, (c) determination of the tracking curve, (d) parasitic effects and distortion due to curvature of valve characteristics, (e) detection. "A work which should not be absent from any technical library."

STATIONS AND COMMUNICATION SYSTEMS

061.3 : 621.396.1 : 621.396.97 992

Results of the Conferences in Copenhagen and Mexico.—E. Metzler. (*Tech. Mitt. schweiz. Telegr.-Teleph.Verw.*, 1st Oct. 1949, Vol. 27, No. 5, pp. 223–231. In German.) Comment on these two conferences, the first on reorganization of European broadcasting channels, the second on international s.w. broadcasting.

621.39.001.11 993

Communication Theory of Secrecy Systems.—C. E. Shannon. (*Bell Syst. tech. J.*, Oct. 1949, Vol. 28, No. 4, pp. 656–715.) Analysis of the basic mathematical structure of secrecy systems and discussion of "theoretical secrecy" and "practical secrecy". See also 1361 of 1949.

621.395 994

15th Plenary Session of the Comité Consultatif International Téléphonique (C.C.I.F.), Paris, July 1949.—R. Sueur. (*Câbles & Transmission, Paris*, Jan. 1950, Vol. 4, No. 1, pp. 3–8.) Summaries of the discussions on various subjects.

621.395.44 : 621.315.052.63 995

Single-Sideband Methods or Frequency Modulation for Power-Line Telephony?—J. Herrmann & J. Erben. (*Frequenz*, Dec. 1949, Vol. 3, No. 12, pp. 341–348.) Discussion of the relative merits of the two systems, with particular reference to corona interference and to distortion due to sideband clipping in f.m. systems.

621.396.41.029.63 : 621.396.97 996

Experimental Ultra-High-Frequency Multiplex Broadcasting System.—A. G. Kandoian & A. M. Levine. (*Elect. Commun.*, Dec. 1949, Vol. 26, No. 4, pp. 292–304.) Reprint. See 2629 of 1949.

621.396.5/.6 997

Description and General Characteristics of [12-channel] Pulse Multiplex Equipment Studied and Perfected at the C.N.E.T. (Centre National d'Étude des Télécommunications).—J. Icole & G. Potier. (*Ann. Télécommun.*, Aug./Sept. 1949, Vol. 4, Nos. 8/9, pp. 313–318.)

621.396.5 : 621.396.619.16 998

Pulse Multiplex Equipment for 24-Channel Telephony.—L. J. Libois. (*Onde élect.*, Jan. 1950, Vol. 29, No. 274, pp. 23–29.) 0.5-μs pulses with a repetition frequency of 8 000 per sec are used. Time distribution of the pulses is effected by means of delay lines. Amplitude modulation of the channel pulses is transformed into pulse-position modulation, a common modulation transformer being used for (*a*) the even-numbered and (*b*) the odd-numbered channels. The transformation procedure is reversed in the receiver. Synchronization pulses are distinguished by their longer duration. Operation of the equipment is satisfactory with a non-stabilized supply. Component tolerances are quite large and though similarity of transmitter and receiver delay lines is important, the precision required does not rule out the possibility of mass production.

621.396.5.029.62 : 621.396.932 999

The Thames Radio Service.—J. Neale & D. W. Burr. (*P.O. elect. Engrs' J.*, Jan. 1950, Vol. 42, Part 4, pp. 213–220.) A description of the recently developed public R/T service for small craft. Full duplex operation is provided. Low-power crystal-controlled transmitters operating in the 160-Mc/s band are used both for the shore-station and the mobile equipments. Mainly because of ease of maintenance and relative cost of equipment, a.m. was adopted for the service in preference to f.m. A selective-calling system is being developed. Details of signalling systems, precautionary measures and maintenance facilities are given. The use of more than one shore transmitter is considered.

621.396.6 : 621.396.932 1000

The Radio Equipment of the Liner "Ile-de-France".—J. Leclère-Courbe. (*Ann. Radioélect.*, Jan. 1950, Vol. 5, No. 19, pp. 21–26.) General description of the functions and disposition of new apparatus which, except for the emergency equipment, is all grouped in one room. It includes magnetostriction and piezoelectric depth-sounding recorders, radar equipment, general communication equipment for both telegraphy and telephony, and radiogoniometer.

621.396.619.13 1001

Quality Gain in Carrier-Frequency Systems using Frequency Modulation of Metre and Decimetre Waves.—P. Barkow. (*Arch. elekt. Übertragung*, Nov. 1949, Vol. 3, No. 8, pp. 287–292.) Formulae are derived for the improvement factors, which are dependent on relative frequency swing and which are represented graphically for different numbers of channels. A numerical example is calculated and two methods are suggested by which equalized signal/noise ratio and additional gain in quality may be obtained, especially for the upper channels.

621.396.619.13 : 621.395.813 1002

Method of Calculating the Nonlinear Distortion of F.M. Signals as a Function of Amplifier Propagation-Time Variations.—J. Fagot. (*Ann. Radioélect.*, Jan. 1950, Vol. 5, No. 19, pp. 8–11.) Phase variation and propagation time of a f.m. signal are defined mathematically and the following simple relation is derived for third-harmonic distortion: $D_3 = \omega\Delta t/4$, where ω is 2π times the modulation frequency and Δt the variation of propagation time. The formula is used to find the maximum permissible variation of t for a given distortion in a 12-channel and a 240-channel carrier-current telephony system.

621.396.619.16 1003

Pulse-Code Modulation and its Distortion with Logarithmic Amplitude Sampling.—H. Holzwarth. (*Arch. elekt. Übertragung*, Nov. 1949, Vol. 3, No. 8, pp. 277–285.) A characteristic of p.c.m. is the increase in distortion with decreasing modulation. This may be partly compensated by a logarithmic pre-distortion of the signals; this acts as a logarithmic sampling of the undistorted signal. Under certain conditions, distortion may be kept at an almost constant value throughout a wide modulation band. An approximate formula is given for the distortion in terms of the degree of pre-distortion and the sampling rate. The method also gives an increased volume range. For the same number of channels, a 7-unit p.c.m. system requires about the same minimum bandwidth as a p.ph.m. system. With a 5-unit code the requirements regarding transmission quality could probably not be fulfilled.

621.396.65 : 621.396.5 1004

A New V.H.F. Multi-Circuit Radio Telephone System: Part I.—J. J. E. Aspin. (*G.E.C. Telecommun.*, 1949, Vol. 4, No. 2, pp. 62–78.) The possibilities of the application of v.h.f. trunk radio systems are summarized, the features of v.h.f. propagation are outlined, and frequency separation is discussed. The effect of sites and aerial heights is explained and figures for attenuation due to the radio path are quoted. An outline of design requirements leads to an illustrated description of equipment of two types: (*a*) the 6-circuit system, which handles seven channels of width 4 kc/s; (*b*) the 12/24-circuit system, which handles 24 such channels. Power output from terminal and relay transmitters is 20 W or, with an additional amplifier, 100 W. Relay spacing may be up to 60 miles. Two rows of 4 centre-fed dipoles, with $\lambda/2$ dipole spacing and λ row spacing, $\lambda/4$ in front of a wire-mesh screen, form the directional

aerial array. Part 2 will describe the automatic changeover and fault-signalling system associated with the duplicate equipment at repeater stations.

621.396.65 : 621.396.619.16 1005

Port Elizabeth/Uitenhage Time-Sharing-Modulation Radio Link.—(*Elect. Commun.*, Dec. 1949, Vol. 26, No. 4, pp. 269–271.) Illustrated description of equipment operating on frequencies of 440 Mc/s and 475 Mc/s in the two directions, with peak power of the pulsed transmitters of 150 W.

621.396.65 : 621.397.5 1006

The London/Birmingham Television Radio-Relay Link.—D. C. Espley & R. J. Clayton. (*G.E.C. J.*, Jan. 1950, Vol. 17, No. 1, pp. 3–10.) An illustrated general description of the present single-channel reversible link. A more detailed technical account will be given on completion of the two-way link. See also 471 of February.

621.396.65(494) 1007

Transalpine Communication by Radio Beams.—W. Gerber. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, 1st Oct. 1949, Vol. 27, No. 5, pp. 231–235. In French.) General discussion, with suggestions for the development of a system linking the regions north and south of the Alps. See also 546 of 1948 (Gerber & Tank) and 2326 of 1949 (Klein).

621.396.822 : 621.396.65 + 621.395.44 1008

Noise Level in F.M. Radio Relay Links for Multichannel Carrier Telephone Systems.—A. van Weel. (*Commun. News*, Oct. 1949, Vol. 10, No. 3, pp. 83–86.) The effective signal/noise ratio of radio links is compared theoretically with that of telephone land lines over the same circuits. It is possible to get comparable performance for the radio link only by using highly directive aerials, involving the use of wavelengths < 30 cm. Economic considerations favour the radio link, since for land lines a repeater is required every 25 km, whereas for the radio link the distance is about 45 km.

621.396.933 1009

Modern Aircraft Radio Equipment as Fitted to Convair Liner.—J. H. Gerrand. (*Proc. Instn Radio Engrs, Aust.*, Nov. 1949, Vol. 10, No. 11, pp. 306–310.) 1948 Australian I.R.E. Convention paper. The three main requirements of such equipment are light weight, reliability and ease of servicing, and simplicity of operation. Various units of the Convair communication and navigation equipment are described, with emphasis on the way in which they meet these requirements.

621.39 1010

Die Systemtheorie der elektrischen Nachrichtenübertragung. (Theory of Electrical Communication Systems) (Book Review)—K. Küpfmüller. Publishers: S. Hirzel, Stuttgart, 386 pp. (*Wireless Engr*, Jan. 1950, Vol. 27, No. 316, p. 30.) " It is a systematic investigation of the transmission of electrical signals of all kinds under all possible conditions. It deals with telegraphic signals, speech, amplitude, frequency and pulse modulation, and all the various types of distortion. The amount of material in the book is very impressive and the treatment is very thorough and detailed and well illustrated with numerous diagrams. When any lesser known mathematical functions are introduced they are carefully explained and illustrated with numerical examples and diagrams. This is certainly a book that can be unreservedly recommended to anyone with a knowledge of German."

SUBSIDIARY APPARATUS

541.136.2 : 541.133 1011

The Internal Resistance and the Electrolytes of Galvanic Cells.—C. Drotschmann. (*Funk u. Ton*, 1949, Vol. 3, Nos. 9/10 & 11/12, pp. 506–524 & 594–600.)

621.316 : 681.142 1012

Electronics and Experimental Mathematics.—Raymond. (*See* 919.)

621.316.722 1013

Photo-Tube Input Impedance for a Voltage Stabiliser.—E. N. Strait & W. W. Buechner. (*Rev. sci. Instrum.*, Nov. 1949, Vol. 20, No. 11, pp. 783–785.) A photocell under constant illumination is used as an input impedance, eliminating the necessity for d.c. amplification of the error signal, which is applied directly to the control grid of a pentode, Type 5693, that is used to control corona leakage current to the h.v. terminal of an e.s. generator.

621.384.5 1014

Variations in the Characteristics of some Glow-Discharge Voltage-Regulator Tubes.—F. A. Benson, W. E. Cain & B. D. Clucas. (*J. sci. Instrum.*, Dec. 1949, Vol. 26, No. 12, pp. 399–401.) " The variations in the characteristics of two commercial types of glow-discharge voltage-regulator tube have been studied in detail experimentally. The results of short- and long-term tests to determine voltage drift, together with measurements of temperature coefficient, are presented. The properties of tubes run with the cathode potential positive with respect to the anode are discussed."

621.396.69 1015

Telescopic Mountings in Electronic Equipment.—C. H. Davis. (*Electronic Engng*, Jan. 1950, Vol. 22, No. 263, pp. 27–31.) A description is given of different methods of mounting chassis on roller bearings or pivots, for easy withdrawal from racks or other equipment and for convenience in servicing. The use of flexible cable harness ensures permanent circuit connections in both open and closed positions. Plug and socket connections are used if the equipment is to be " dead " in the open position.

TELEVISION AND PHOTOTELEGRAPHY

621.397.5 1016

1029-Line Television.—W. Dillenburger. (*Elektron Wiss. Tech.*, Jan. 1950, Vol. 4, No. 1, pp. 1–10.) The suitability of different scanning methods for high-definition pictures is discussed and their efficiency considered. Tests made with a 1029-line image dissector tube indicate that this line number is too high and that the physical limit for scanning tubes is too closely approached. The cost of transmitters and receivers is considered. The proposed 625-line system for German television seems a good choice, guaranteeing satisfactory picture quality and being economical.

621.397.5 : 621.396.65 1017

The London/Birmingham Television Radio-Relay Link.—Espley & Clayton. (*See* 1006.)

621.397.62 1018

Technique and Developments of High-Definition Television Receivers.—P. Mandel. (*Onde élect.*, Jan. 1950, Vol. 29, No. 274, pp. 45–46.) Paper presented at the 1949 International Television Congress, Milan. An examination of problems of technique and economy. The technical aspects discussed are: aerials and h.f. feeders; h.f. amplification and frequency-changing; i.f. amplifiers and detection; electron beam deflection and image distortion. The relative cost of high- and low-definition receivers is considered and the importance of the standardization of components is stressed.

621.397.7 1019

Extending Television.—(*Wireless World*, Jan. 1950, Vol. 56, No. 1, pp. 14–15.) Brief description of Sutton Coldfield television station, which operates at 61.7 Mc/s, with a power of 35 kW, and is the second of a chain of 10 stations which are planned for extending the service to 80% of the population of Great Britain by 1954.

TRANSMISSION

621.39.001.11 1020

Analytical Signals with Limited Spectrum: Part 1.—J. Ville. (*Câbles & Transmission, Paris*, Jan. 1950, Vol. 4, No. 1, pp. 9–23.) The properties of signals whose spectra lie within a given frequency band are studied mathematically. Such signals are completely defined by their values at instants in arithmetical progression with a sufficiently small common difference. The extension of this result to analytical signals permits the study of circuits whose indicial admittance is distributed in time. This enables the relation between the bandwidth and the quantity of information carried by the signal to be determined.

621.396.612.1.029.64 1021

A New Transmitter for Ultra-Short Damped Waves.—J. Le Bot. (*J. Phys. Radium*, May 1947, Vol. 8, No. 5, pp. 142–145.) Description of a Hertzian doublet type of oscillator of simple construction. Two tungsten or platinum wires, of diameter 1 mm and length 4.5 mm, are arranged as a $\lambda/2$ doublet with a gap of some hundredths of a millimetre which has a micrometer adjustment. The doublet is immersed in a mineral oil and fed through fine resistance wires connected through a high resistance to an 8–10-kV a.c. supply. At breakdown the spark discharge carbonizes the oil to form an effective short circuit for the h.f. oscillation generated. The resistive component of the impedance of the doublet being very small, a damped wave-train is radiated at each succeeding half-cycle. The apparatus functions for several hours without adjustment. Radiated power is of the order of 10–20 mW and wavelength about 4.4 cm. Some experiments with cm waves are described, using this transmitter coupled to a short waveguide and horn radiator.

621.396.615 : 621.396.619.13 1022

Reactance-Tube Modulation of Phase-Shift Oscillators.—F. R. Dennis & E. P. Felch. (*Bell Syst. tech. J.*, Oct. 1949, Vol. 28, No. 4, pp. 601–607.) A basic circuit is described and the design of suitable oscillators for a range of frequencies from a.f. to u.h.f. is discussed. The phase-shift networks are conveniently of the RC, LC, and transmission-line types for the lower, medium and higher frequencies respectively. Typical circuits and performance curves are given.

621.396.619.13 1023

Calculation of the Bandwidth of a Sinusoidal Transmission with Sinusoidal F.M.—L. Robin. (*Ann. Télécommun.*, Jan. 1949, Vol. 4, No. 1, pp. 19–26.) The bandwidth necessary to satisfy the requirements fixed at the Atlantic City Convention is determined for any value of the modulation index. The treatment involves Bessel functions of the first kind and a series formed by the squares of these functions. It is found necessary to introduce an asymptotic development of Bessel functions for neighbouring values of their index and argument. The results of the calculations are shown on four graphs. Two of these give the ratio of f.m. bandwidth to a.m. bandwidth for any value of the modulation index; the others show the relation between the f.m. bandwidth and the frequency swing.

621.396.619.13 1024

Reactance Valves and their Use in Frequency-Modulation Circuits.—A. G. Wray. (*Marconi Instrumentation*, Nov./Dec. 1949, Vol. 2, No. 4, pp. 63–67.) Simple theory is given of the operation of a typical reactance-valve circuit. This indicates that a valve giving a linear change of mutual conductance with linear variation of control-grid voltage is suitable for use as a reactor. Experiments with a variable-μ valve showed a reduction in second-harmonic content but an increase of the third harmonic, so that no improvement of the total distortion resulted. With a 'straight' valve the third harmonic is practically non-existent and by careful adjustment of bias voltage the second harmonic can also be made negligible for a particular valve. Bias readjustment will probably be necessary if the valve is changed for one of the same type. Using an EF91, EL91, or EF92 valve in the circuit given, a deviation of 0.5% of the carrier frequency can be obtained. Much better performance is obtained with a push-pull circuit, full details of which are given. A deviation of 2% of the carrier frequency is possible with this circuit, using EF91 valves for both reactor and oscillator. Overall envelope distortion is $< 1\%$. The system has the advantage that random changes due to supply-voltage fluctuations are cancelled.

621.396.619.23 1025

Study of Rectifier-Type Ring Modulators with regard to their Application in Terminal Installations for 600- or 960-Channel Coaxial-Cable Links.—P. Moll. (*Câbles & Transmission, Paris*, Jan. 1950, Vol. 4, No. 1, pp. 24–46.) The characteristics of Westinghouse rectifiers 30S9A and 28S4F and their associated input and output transformers are studied at different frequencies between 10 kc/s and 10 Mc/s. Laboratory tests and methods of measurement of circuit losses and nonlinear distortion are described. Results are tabulated and shown graphically. The 30S9A rectifier is well suited for ring-modulator use between 300 c/s and 4 Mc/s.

VALVES AND THERMIONICS

537.311.33 : 621.315.59 1026

Transistors.—J. A. Becker. (*Elect. Engng, N.Y.*, Jan. 1950, Vol. 69, No. 1, pp. 58–64.) Discussion of recent advances in the theory, circuit performance and potential applications of transistors, based on the following papers presented at the 1949 A.I.E.E. Summer Meeting:

Theory of Transistor Action.—W. Shockley.

Equivalent Circuits for Transistor Action and Noise.—R. M. Ryder.

The Possible Significance of Transistors in the Power Field.—J. A. Hutcheson.

See also 240 of January.

621.383.4 : 534.862.4 1027

Lead-Sulfide Photoconductive Cells in Sound Reproducers.—R. W. Lee. (*J. Soc. Mot. Pict. Engrs*, Dec. 1949, Vol. 53, No. 6, pp. 691–703. Discussion, pp. 703–706.) Results are given of measurements of sensitivity or signal output for a wide range of conditions in a standard 16-mm sound projector, using silver, silver-sulphide, dye-image, and iron-toned sound tracks. The effect of the colour temperature of the exciter lamp on frequency response and signal output is also discussed.

621.384.5 1028

Variations in the Characteristics of some Glow-Discharge Voltage-Regulator Tubes.—Benson, Cain & Clucas. (*See* 1014.)

621.385 : 621.318.572 1029

Switching Valves.—W. Kleen. (*Elektron Wiss. Tech.*, Jan. 1950, Vol. 4, No. 1, pp. 11–16.) The principle of operation, application and development of the gas-filled transmit-receive switch.

621.385 : 621.396.621 1030

Modern Receiving Valves: Miniature Valves.—R. Suart. (*Radio franç.*, Jan. 1950, No. 1, pp. 3–8.) An illustrated description of their development and of methods of manufacture. Operating characteristics compare favourably with those of the corresponding normal and acorn types of valve. Some 40 types of miniature valve are classified.

1031

621.385 : 621.396.621

The Sensitivity of Receiving Valves.—Rothe. (*See* 986.)

1032

621.385.029.63 : 64 : 537.525.92

Increasing Space-Charge Waves.—Pierce. (*See* 874.)

1033

621.396.615.029.63

On a Particular Characteristic of Disk-Seal Valves 2C40 and 2C43.—L. Liot. (*Radio franç.*, Feb. 1950, No. 2, pp. 9–14.) The physical construction of these valves allows them to be used as reentrant-cavity oscillators for the frequency range 1 600–2 400 Mc/s. Maximum output power using a Type-2C40 valve on 2 000 Mc/s is 200 mW, falling to 70 mW at the limiting frequencies. A scale diagram is given showing the construction adopted.

1034

621.396.615.142

On the Theory of Velocity-Modulation Transit-Time Valves: Part 2.—H. Döring. (*Arch. elekt. Übertragung*, Nov. 1949, Vol. 3, No. 8, pp. 293–300.) The general reflex type of valve is investigated. Behind the infinitely short alternating field is located the field-free drift space θ_L and a uniform retarding field θ_R of finite length. This arrangement is comparable with the two-field, two-circuit valve with fixed reaction, so that discrete oscillation regions appear when the static transit-time angle is varied. When $\theta_L = \theta_R$ the oscillation regions correspond to those of the Heil generator with antiphase alternating fields. When $\theta_L < \theta_R$ the motion in the retarding field is predominant and the oscillation regions coincide with the damping regions of the Heil generator. For equal transit-time angle in the drift space and retarding field ($\theta_L = \theta_R$), only slight density modulation occurs and the efficiency of the generator tends to zero. Two arrangements giving the greatest conversion efficiency are described. The first uses an electrical double layer for the retarding field ($\theta_R = 0$). In this case, oscillation regions and efficiency are identical with those of the Heil generator. Maximum efficiency of 26% occurs in the second region with $\beta = 0.5$ and $\theta_L = 2.3\pi$. In the second arrangement $\theta_L = 0$ and the maximum efficiency of 25.5% occurs in the first region with $\theta_R = 1.55\pi$ and $\beta = 0.5$. Exact calculations are made for the first and sixth regions. Approximate formulae may be derived for the maximum efficiency in the higher oscillation regions. Part 1: 2175 of 1944.

1035

621.396.822

Spontaneous Fluctuations in Double-Cathode Valves.—R. Fürth. (*Wireless Engr*, Jan. 1950, Vol. 27, No. 316, p. 33.) Reply to comment by Knol & Diemer (791 of March) on Fürth's theory (2418 and 2419 of 1948). See also 1247 of 1949 (MacDonald).

1036

621.396.822

Hum in A.C. Valves.—Z. Imre. (*Electronic Engng*, Jan. 1950, Vol. 22, No. 263, p. 33). See also 916 of 1949, where the author's name should be as above.

1037

621.38

Fundamental Electronics and Vacuum Tubes. [Book Review]—A. L. Albert. Publishers: Macmillan & Co., 2nd edn, 1947, 499 pp., $6.00. (*Proc. Inst. Radio Engrs, W. & E.*, Nov. 1949, Vol. 37, No. 11, pp. 1304–1305.) Intended "for use in courses in electronics required of all electrical engineering students. The treatment is largely non-mathematical." The first edition was noted in 2783 of 1939.

1038

621.385

International Radio Tube Encyclopedia. [Book Review]—B. R. Babani (Ed.). Publishers: Bernards Ltd, London, 410 pp., 42s. (*Wireless Engr*, Dec. 1949, Vol. 26, No. 315, p. 412.) Gives the characteristics and base connections of 15 000 British, American and European valves, classified according to type and arranged in alphabetical and numerical order.

1039

621.385

Les Tubes Électroniques et leurs Applications: Tome 1 — Principes généraux. [Book Review]—H. Barkhausen. Publishers: Dunod, Paris, 228 pp. (*Radio tech. Dig., Édn franç.*, Dec. 1949, Vol. 3, No. 6, pp. 377–378.) 2nd French edition, being a translation by C. Poitrat of the 5th German edition. "This book, like the earlier edition, will be of great value to radio engineers and technicians." Further volumes will deal with amplifiers (vol. 2), self-excitation and feedback (vol. 3), and rectifiers and receivers (vol. 4).

MISCELLANEOUS

1040

06.064 : 621.396

Amateur Exhibition.—(*Wireless World*, Jan. 1950, Vol. 56, No. 1, pp. 12–13.) Brief descriptions of some of the exhibits at the 3rd annual show of the Radio Society of Great Britain.

1041

530.12

Albert Einstein on His Seventieth Birthday.—R. A. Millikan. (*Rev. mod. Phys.*, July 1949, Vol. 21, No. 3, pp. 343–345.) A short appreciation of Einstein and his work.

The rest of this number (pp. 345–540) contains 37 papers on various aspects of the theory of relativity. Each of the authors would wish to express his debt to Einstein, but editorial arrangements made it necessary for such expressions to be voiced only by R. A. Millikan, L. de Broglie, M. von Laue and P. Frank.

1042

621.3 : 371.3

Philosophy of the Teaching of Electricity.—L. Bouthillon. (*Bull. Soc. franç. Elect.*, Nov. 1949, Vol. 9, No. 98, pp. 619–640.)

1043

621.385

Birth Centenary of Ambrose Fleming.—J. T. MacGregor-Morris. (*Electronic Engng*, Dec. 1949, Vol. 21, No. 262, pp. 442–447 & Jan. 1950, Vol. 22, No. 263, pp. 22–26.) A review of Fleming's life and work, with particular reference to the sequence of events leading to his invention of the thermionic valve and to its use for radio reception.

1044

621.396.029.64

Some Experimental Demonstrations with Electromagnetic Centimetre Waves.—E. Meyer & H. Severin. (*Z. Phys.*, 30th Aug. 1949, Vol. 126, Nos. 7/9, pp. 711–720.) Description of simple experiments in an auditorium of Göttingen University during the course of a lecture on "Physical Fundamentals of High-Frequency Technique". A 10-cm tone-modulated magnetron was used as transmitter.

1045

621.396.97 : 061.3

Proceedings of the Broadcast Engineers Conference, May 23-25, 1949.—M. A. Honnell (Ed.). (*Georgia Inst. Technol. State Engng Exp. Station Circular, Atlanta*, Oct. 1949, Vol. 9, No. 8, 82 pp.) Essential substance of 14 papers read at the conference.

1046

43-3 = 2 = 4 : 621.3

Technical Dictionary: Vol. 2, German-English-French. [Book Review]—H. Thali. Publishers: H. Thali & Co., Hitzkirch, Switzerland, 1948, 311 pp., 35s. (*Electronic Engng*, Nov. 1949, Vol. 21, No. 261, p. 437.) A 24 000-word dictionary concerned mainly with electrical engineering, radio, television and telecommunications. It includes also the most important terms of acoustics, illumination, mathematics, materials, optics, heating, etc. Vol. 1, English-German-French, was noted in 906 of 1948.

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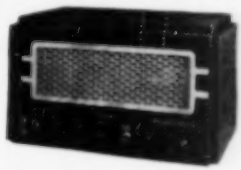

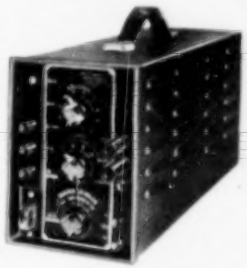

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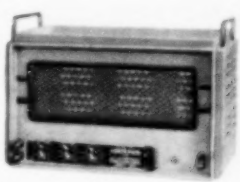

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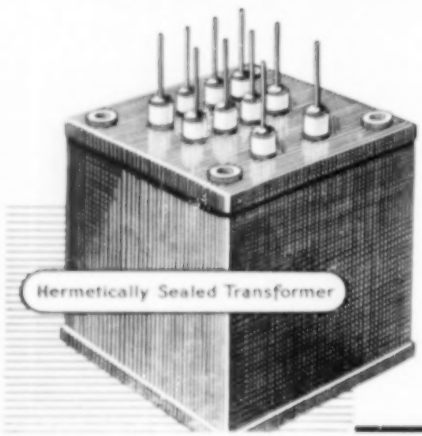

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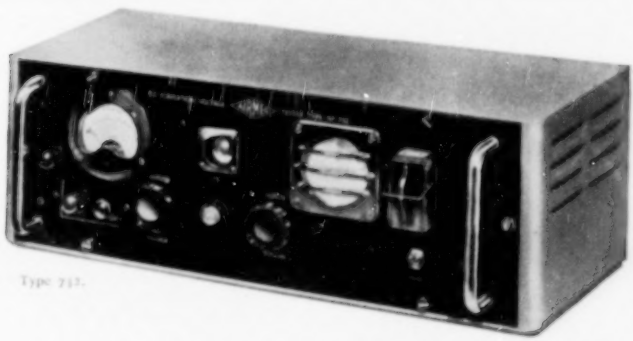

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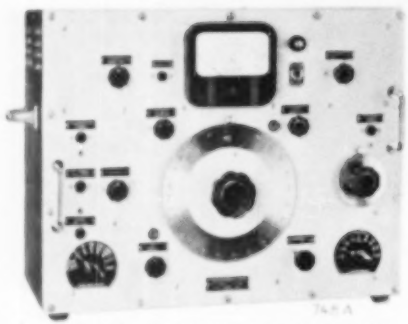

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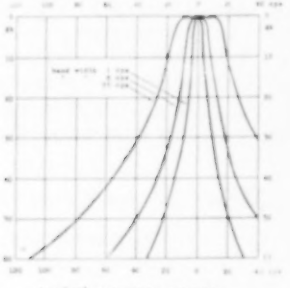

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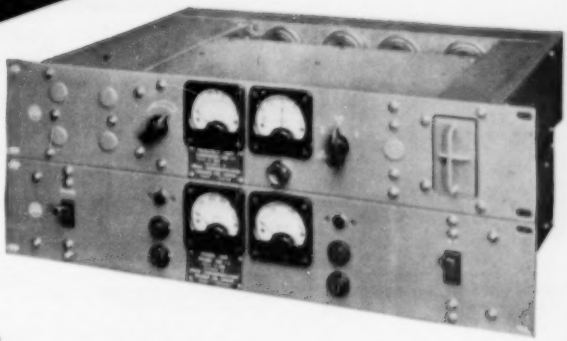

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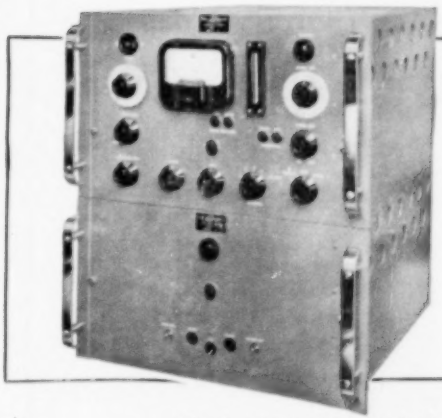

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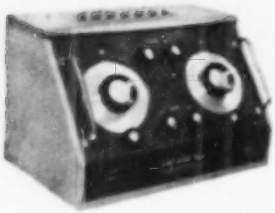

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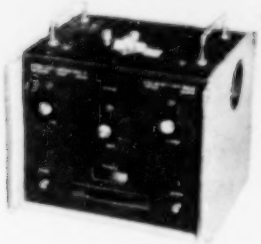

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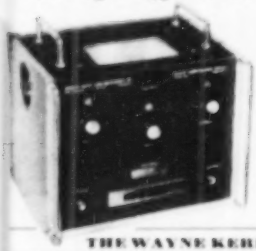

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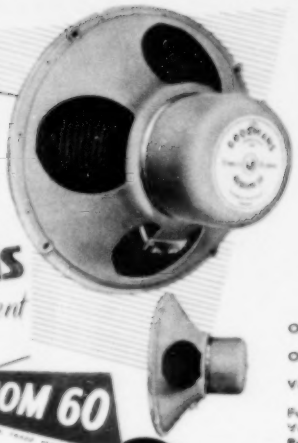

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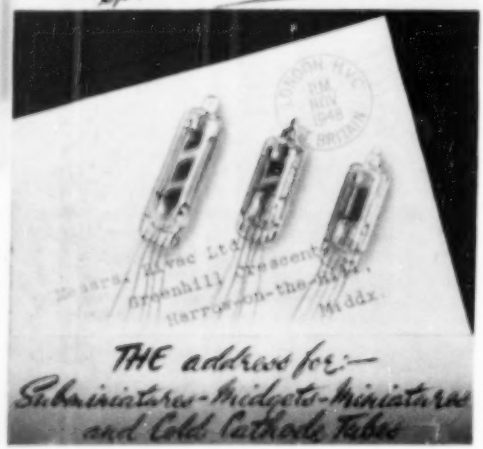

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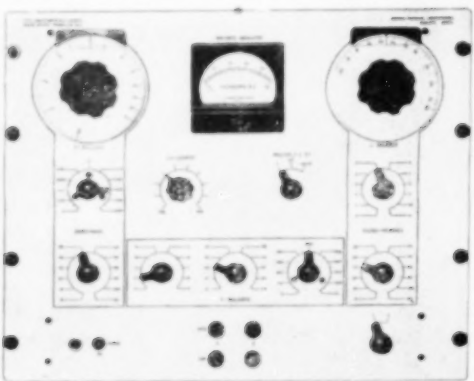

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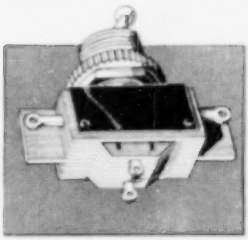

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.002	500	350	1 in.	¼in.	CP30S
.01	500	350	1 in.	3/16in.	CP33S
.05	500	350	1¼in.	¼in.	CP37S
.005	350	200	1 in.	7/32in.	CP31N
.02	350	200	1 in.	3/16in.	CP33N
.1	350	200	1¼in.	¼in.	CP37N

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.002	350	200	¼in.	0.22in.	CP111N
.01	200	120	¼in.	0.25in.	CP112N
.01	350	200	¼in.	0.34in.	CP113N

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For those who were unable to meet us at our Stand No. 57 at the R.E.C.M.F. Grosvenor House Exhibition we show below one of our advertisement pages from the Catalogue.

WHY Ersin Multicore IS

The Finest Cored Solder in the World

Three cores of flux ensure flux continuity throughout the length of the solder wire. There are no lengths without flux—that means no wasted solder, no wasted time and freedom from 'dry' or H.R. joints.

Soldered joints made with Ersin Flux do not corrode even after prolonged exposure to any degree of humidity. This has been tested under climatic conditions ranging from the Arctic to the Tropics.

It is more economical to use the solder with the smallest practicable flux percentage. The extra activity of Ersin Flux and the rapid melting—due to the Multicore construction—enables the Ersin Multicore Solder with 2.2% flux content, to be used for most assembly soldering processes. This is more than 1% less flux content than most other solders.

The correct proportions of flux to solder are always assured—no extra flux is required. Only by having three separate cores of flux can instantaneous melting be obtained.

ALLOYS. This table shows the Melting Points of the three principal alloys used in Electronic and Telecommunication industries.

Alloy	Colour Code	Sol.	Liq.	Bit. Temp.	USES
60/40	Red	183°C	190°C	230°C	High quality work
45/55	Crimson/Buff	183°C	227°C	267°C	Hand soldering radio and electrical
40/60	Green	183°C	238°C	278°C	

GAUGES. Approximate number of feet per lb. (3.4% flux content)

Standard Wire Gauge	Diam. in inches	Diam. in M.ms.	ALLOY 60/40	45/55	40/60
10	0.128	3.251	25.2	23.5	23.0
12	0.104	2.642	38.1	35.2	34.9
13	0.092	2.337	48.7	45.3	44.5
14	0.080	2.032	64.4	59.2	58.6
16	0.064	1.626	100.5	94.3	92.1
18	0.048	1.219	178.5	167.6	163.5
19	0.040	1.016	257.5	240.4	235.5
20	0.036	0.914	318.0	302.5	291.0
22	0.028	0.711	526.0	492.0	481.0

Ersin Multicore Solder is made as standard for factory use in 5 alloys and 9 gauges, and is supplied on nominal 7 lb. reels. Other alloys and gauges can be supplied to special order. Bulk prices on application.

The economies effected by using Ersin Multicore Solder play an important part in cutting production costs. You get more joints per lb. of Ersin Multicore — there is no waste. Soldering with Ersin Multicore is quicker, and you know that you are using a product that has given world wide satisfaction for more than 10 years.

U.S.A. IMPORTS

BRITISH MADE ERSIN MULTICORE SOLDER

Ersin Multicore Solder is so much more reliable and efficient than any other cored solder that even many leading U.S.A. electronic manufacturers import Ersin Multicore despite freight and import duties.

(Right) Assembling television receivers at the Du Mont Factories, Newark, U.S.A. with British made Ersin Multicore Solder.

Multicore Solders and Ersin Flux are entirely of British invention, formulation and manufacture.

ERSIN FLUX

Ersin, the flux contained in the three cores of Ersin Multicore, is a high grade rosin which has been subjected to a complex chemical process to increase its fluxing action, without impairing the well-known non-corrosive properties of the original rosin

We will be pleased to send to Laboratory Engineers and Technicians, new technical literature containing useful tables of melting points, etc., and samples of alloys.

MULTICORE SOLDERS LTD.

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